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Ideal diode controllers

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Audio Out
Scratchy potentiometers: analysis and solutions

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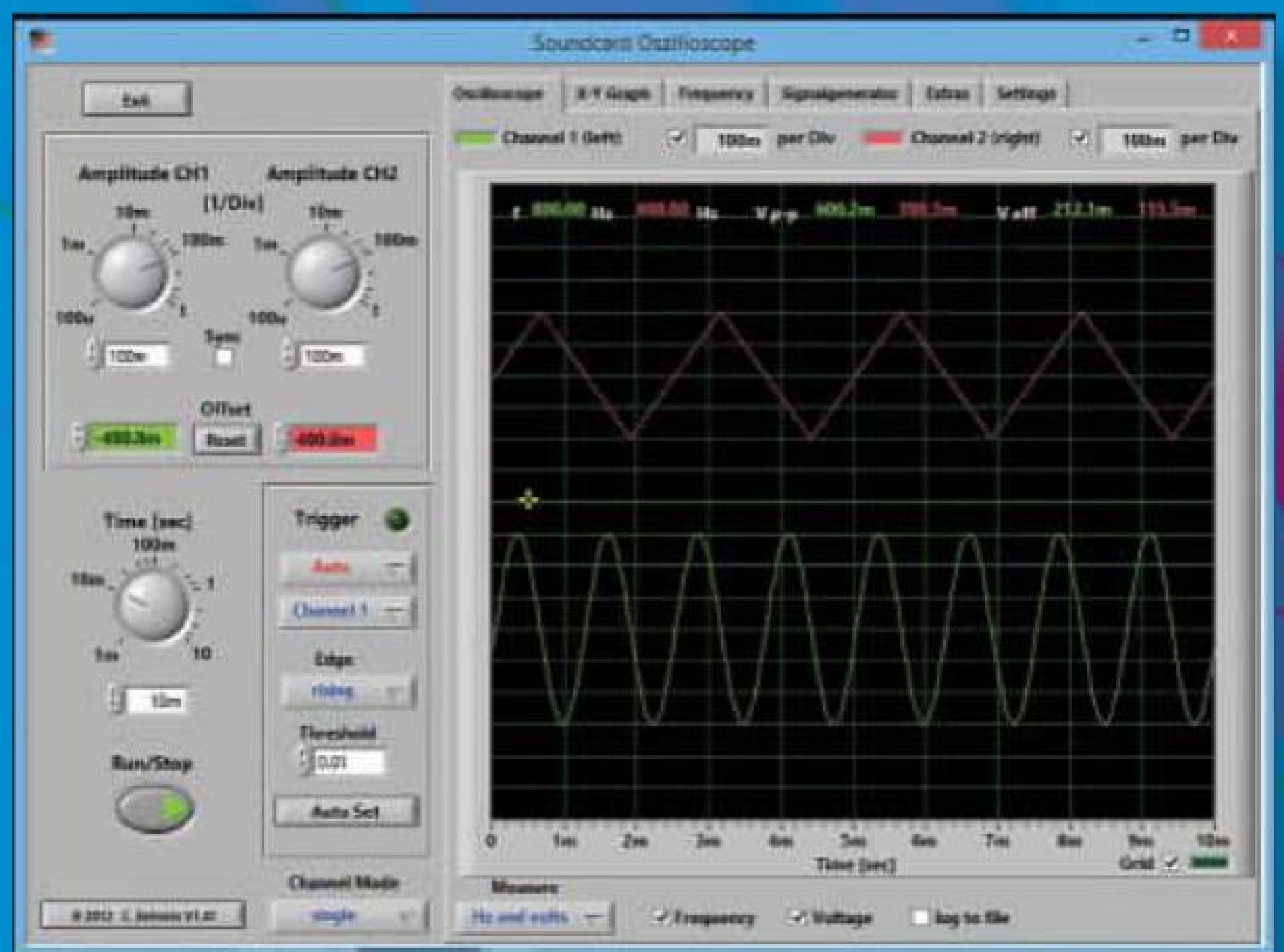

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Choosing your first 'scope

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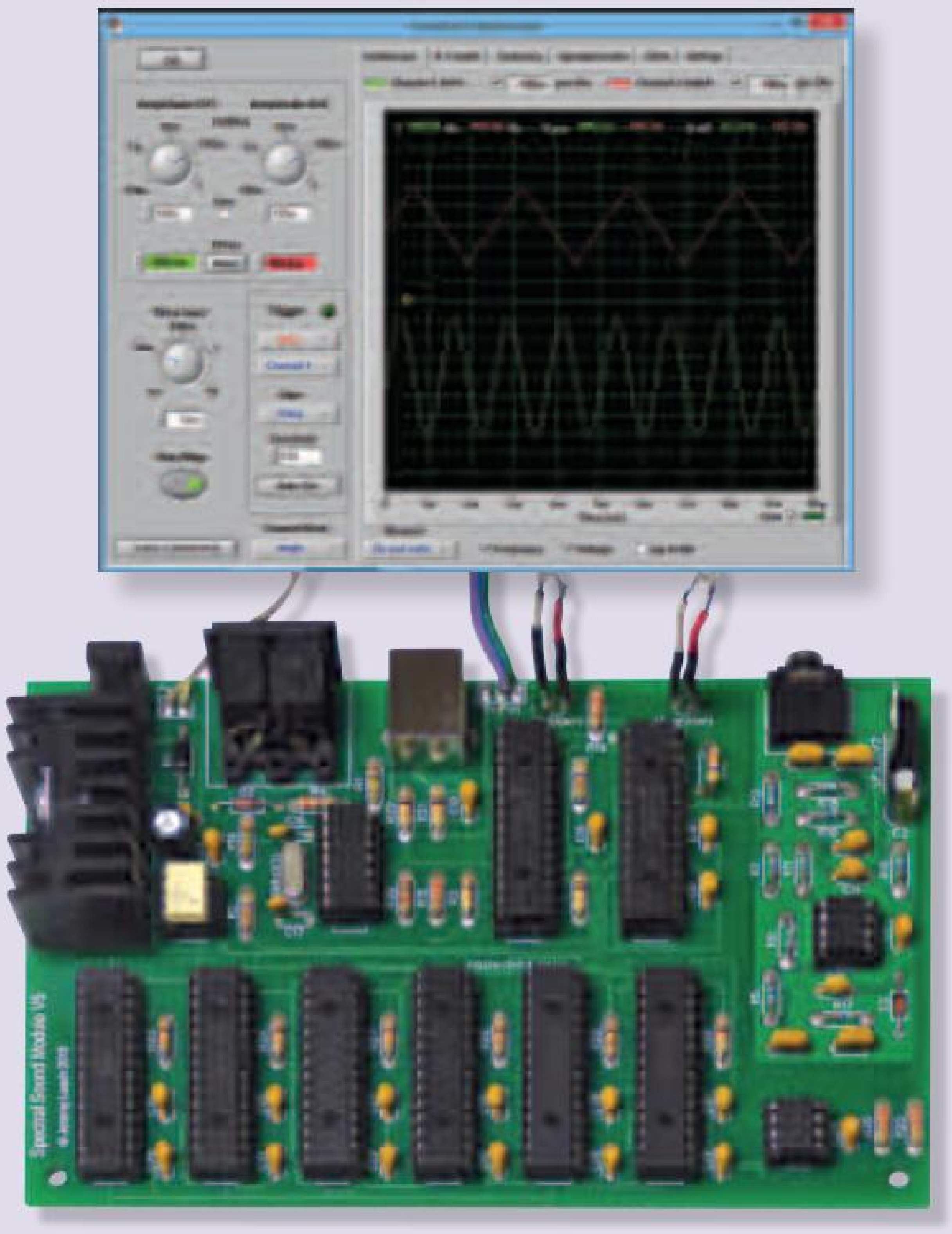

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The August 2023 issue of *Practical Electronics* will be published on Thursday, 6 July 2023 – see page 72.

WIRELESS FOR THE WARRIOR

by LOUIS MEULSTEE

THE DEFINITIVE TECHNICAL HISTORY OF RADIO COMMUNICATION EQUIPMENT IN THE BRITISH ARMY

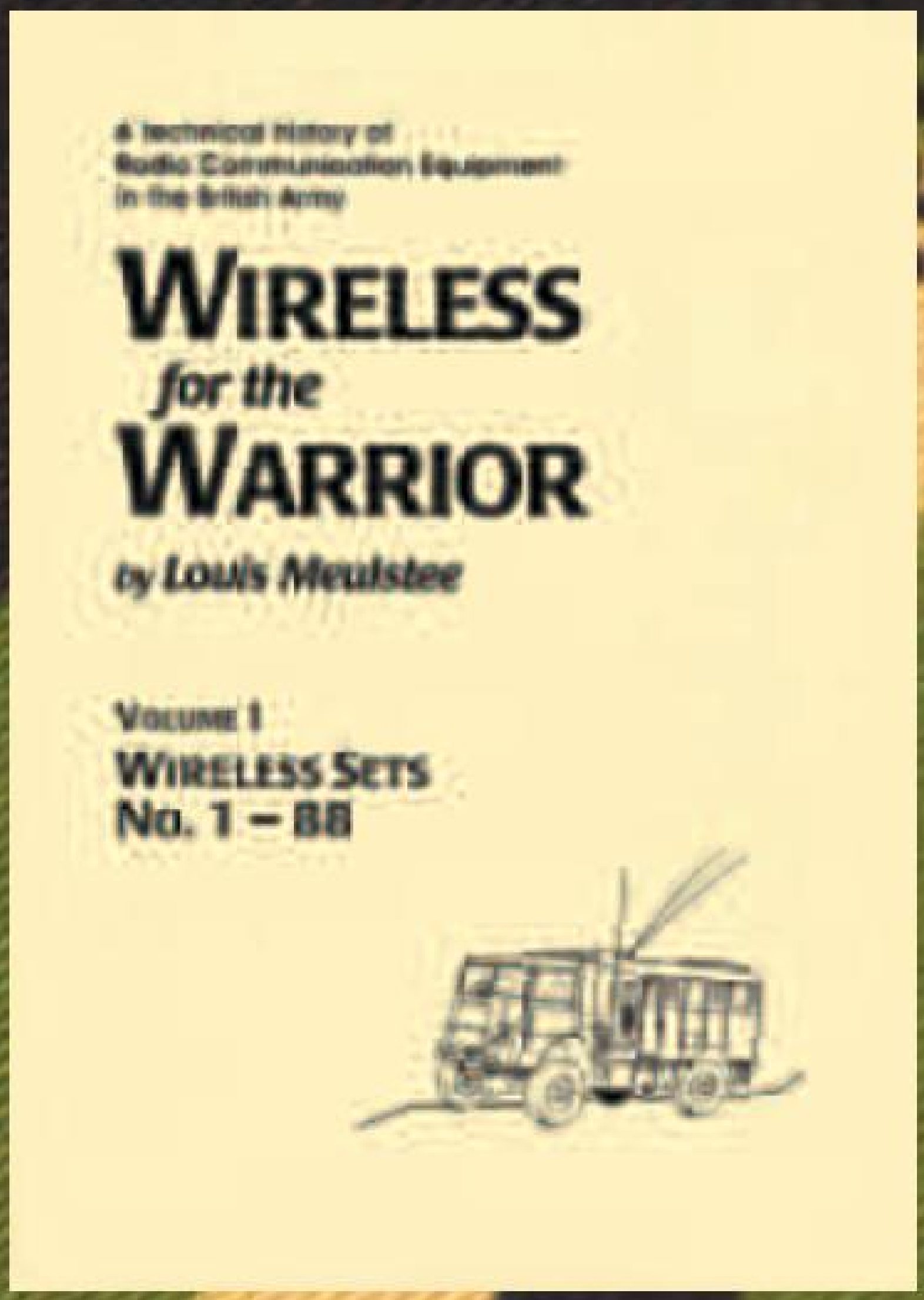
The *Wireless for the Warrior* books are a source of reference for the history and development of radio communication equipment used by the British Army from the very early days of wireless up to the 1960s.

The books are very detailed and include circuit diagrams, technical specifications and alignment data, technical development history, complete station lists and vehicle fitting instructions.

Volume 1 and *Volume 2* cover transmitters and transceivers used between 1932-1948. An era that starts with positive steps taken to formulate and develop a new series of wireless sets that offered great improvements over obsolete World War I pattern equipment. The other end of this timeframe saw the introduction of VHF FM and hermetically sealed equipment.

Volume 3 covers army receivers from 1932 to the late 1960s. The book not only describes receivers specifically designed for the British Army, but also the Royal Navy and RAF. Also covered: special receivers, direction finding receivers, Canadian and Australian Army receivers, commercial receivers adopted by the Army, and Army Welfare broadcast receivers.

Volume 4 covers clandestine, agent or 'spy' radio equipment, sets which were used by special forces, partisans, resistance, 'stay behind' organisations, Australian Coast Watchers and the diplomatic service. Plus, selected associated power sources, RDF and intercept receivers, bugs and radar beacons.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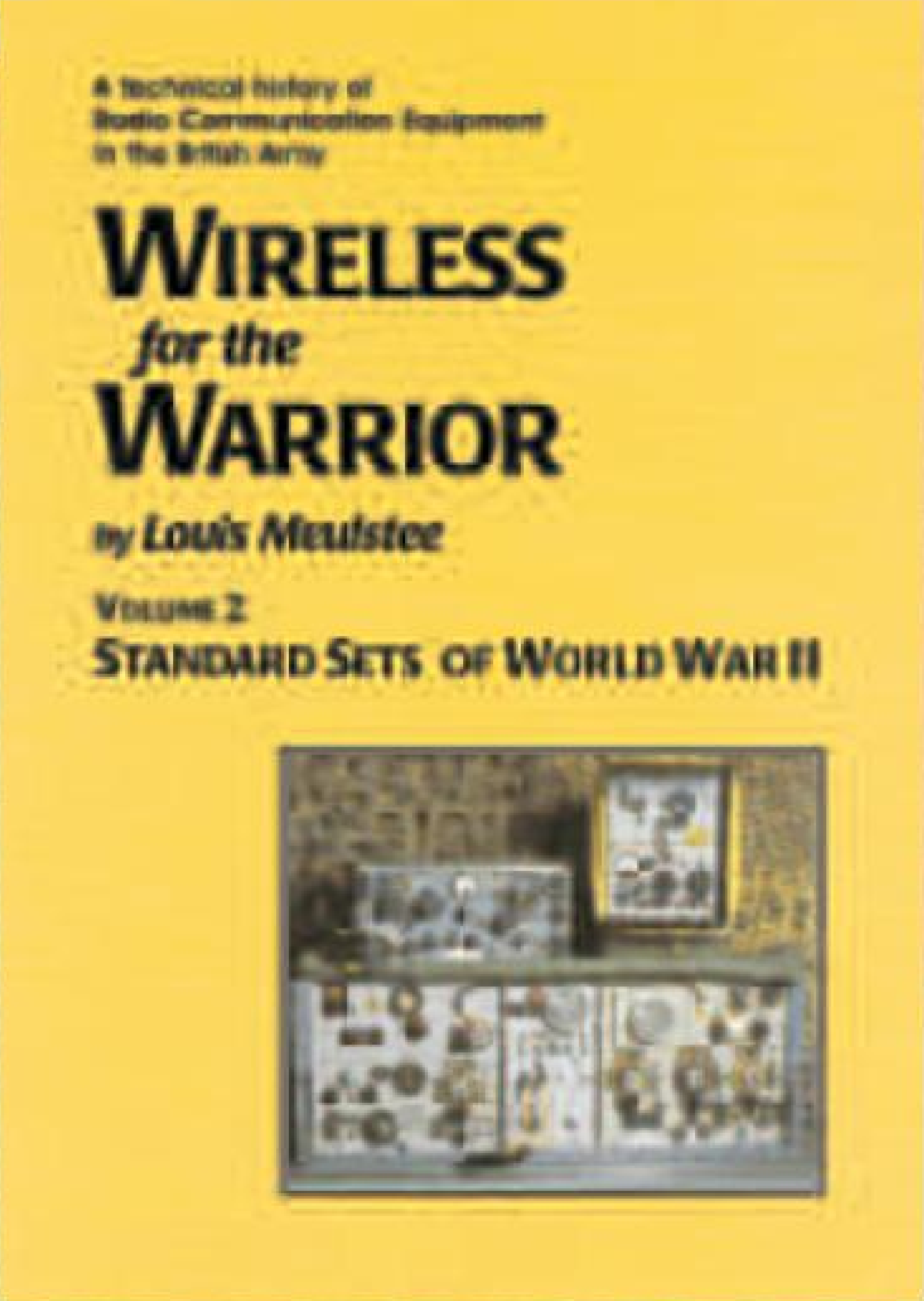
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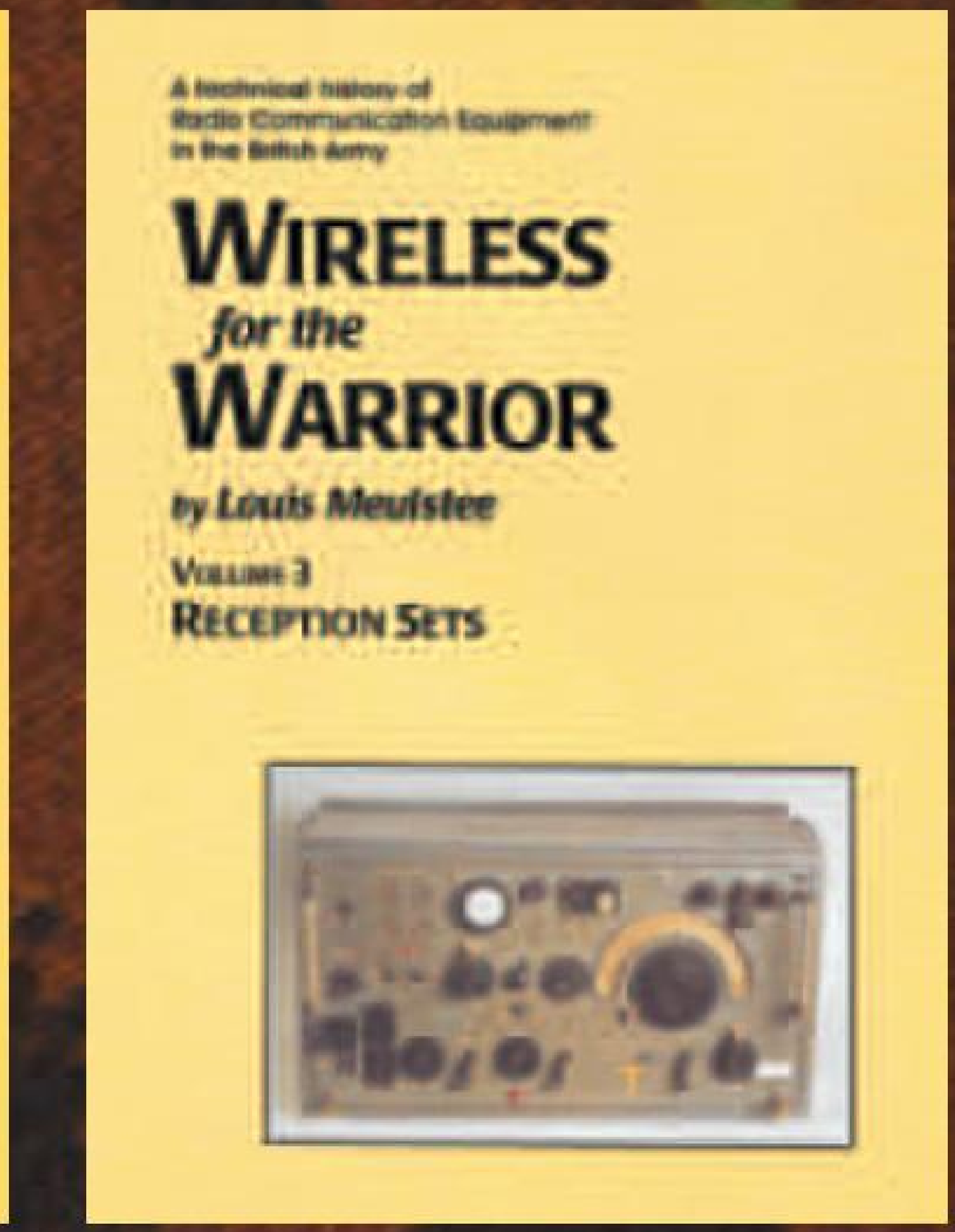
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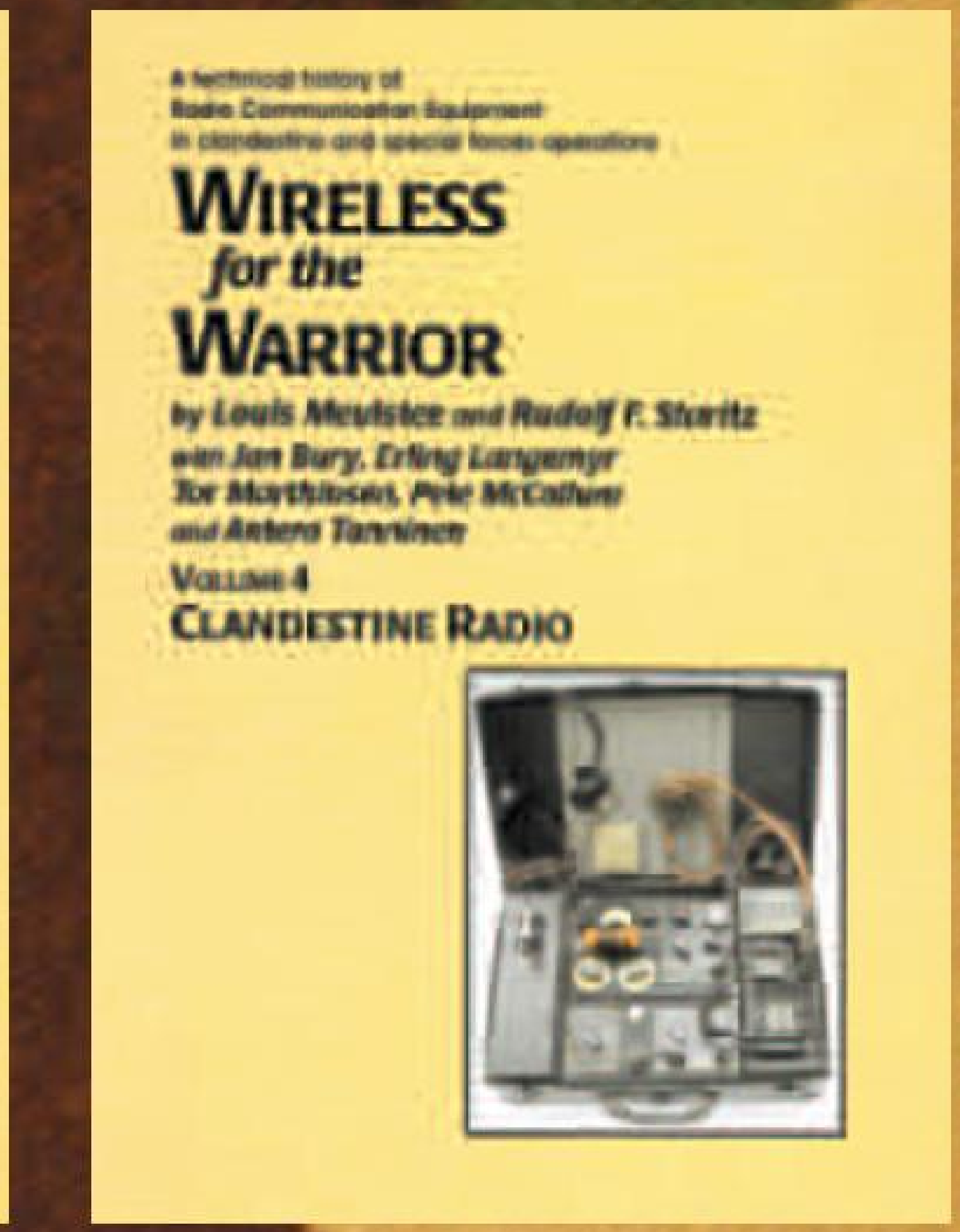
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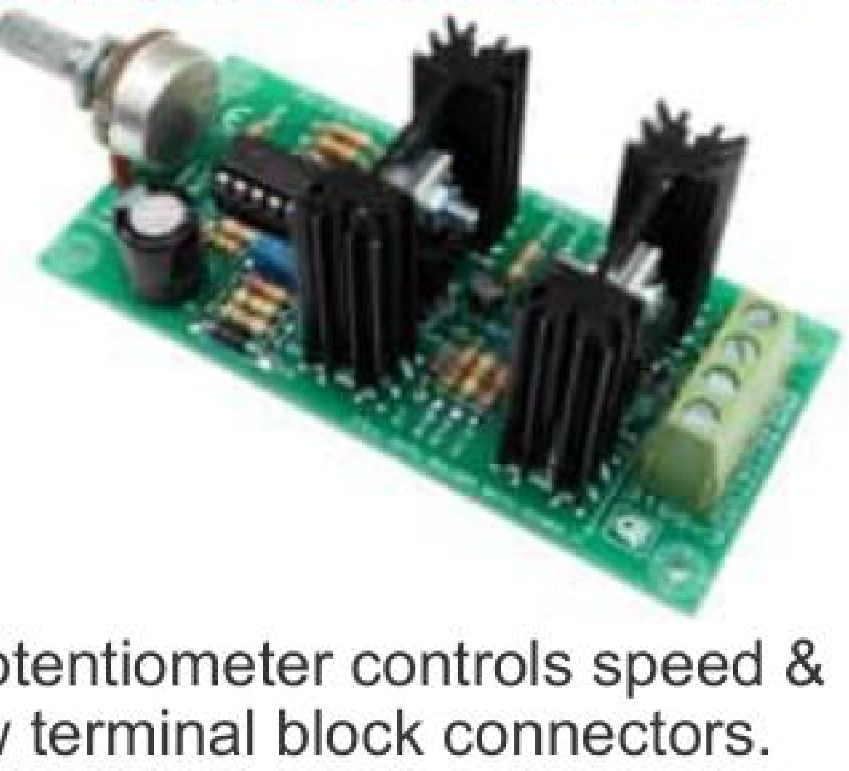

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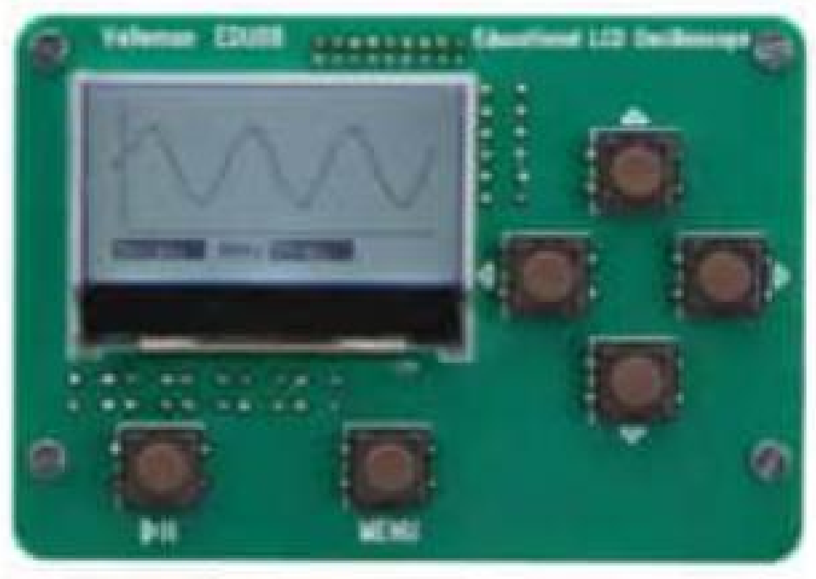

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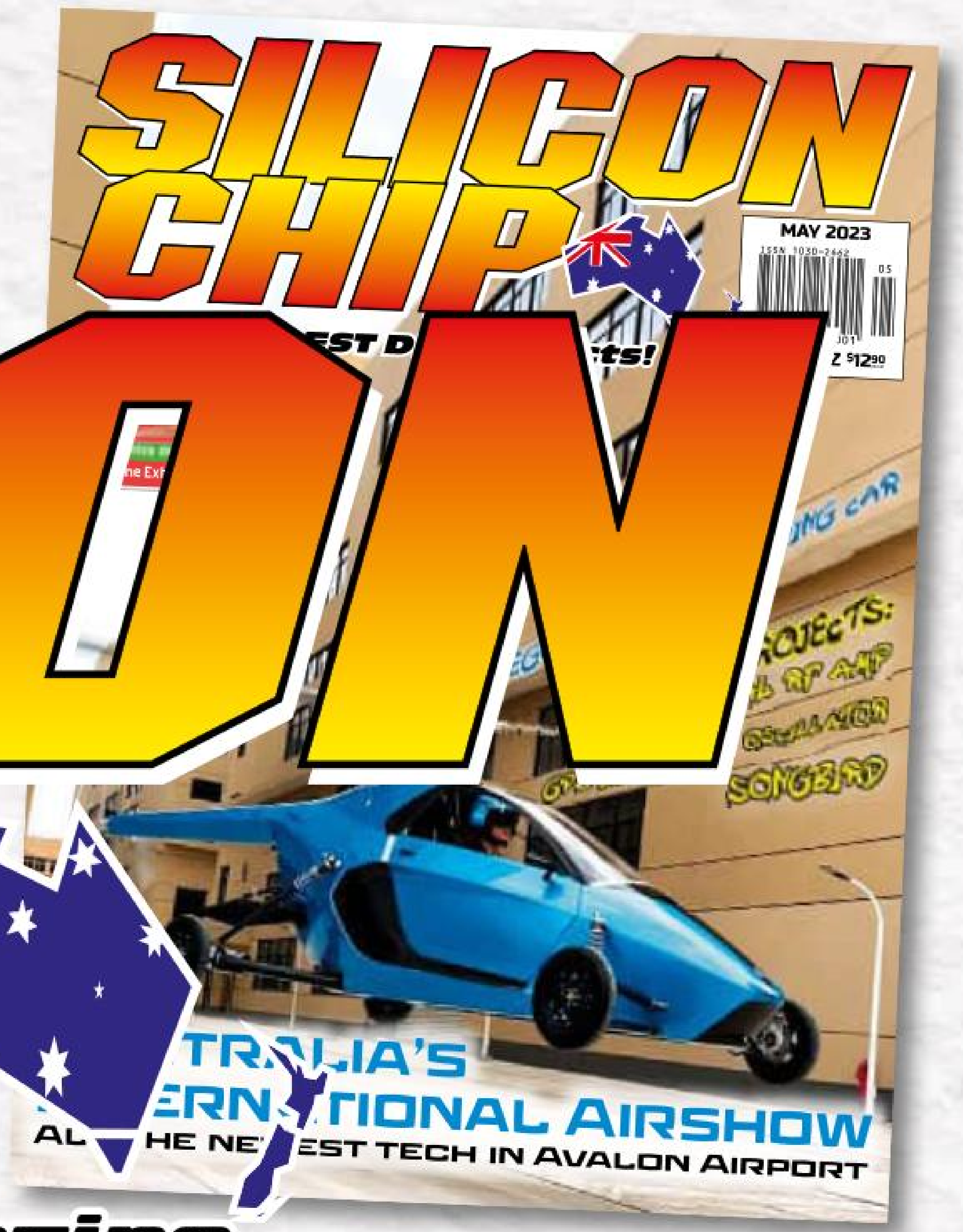
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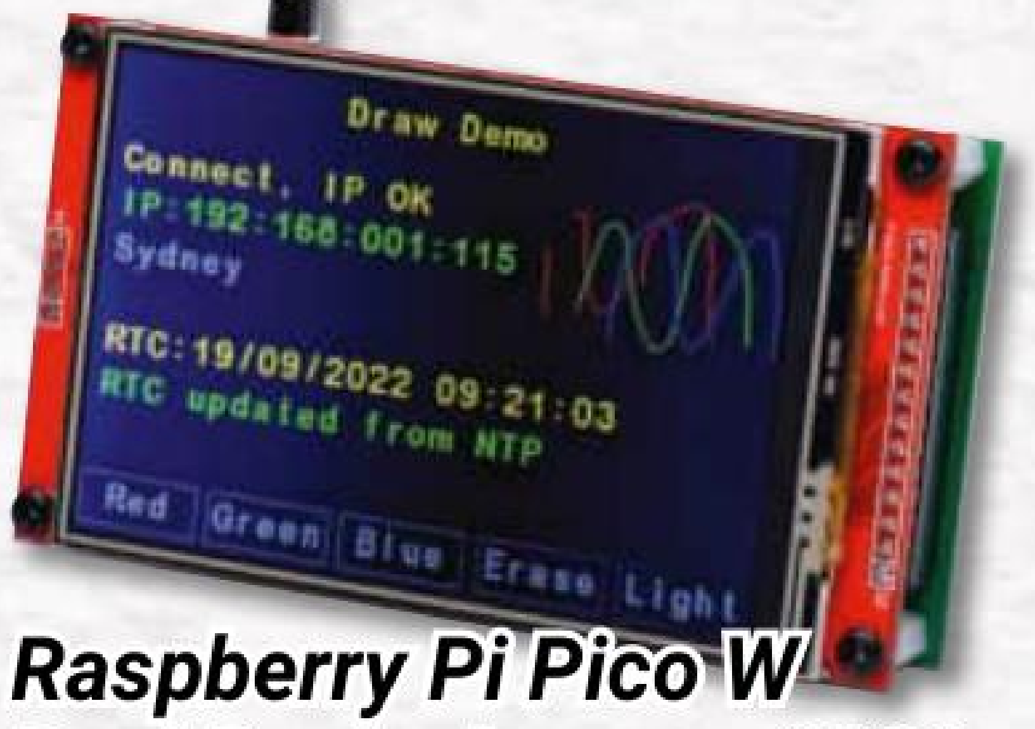
Draw Demo
Connect, IP OK
IP:192:168:001:115
Sydney
RTC:19/09/2022 09:21:03
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A number of projects and circuits published in *Practical Electronics* employ voltages that can be lethal. You should not build, test, modify or renovate any item of mains-powered equipment unless you fully understand the safety aspects involved and you use an RCD (GFCI) adaptor.

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Check your meter

Last month, Julian Edgar wrote an excellent review of the EEVBlog 121GW multimeter. It's a very nice product, but perhaps not one all of us could afford. Nevertheless, we all need a meter of one kind or another, and these days it is almost always going to be a DVM (digital voltmeter), which despite its name will likely include the ability to measure AC/DC currents and resistance. (Capacitance, frequency and diode parameters are often part of the mix too.)

So, perhaps you settle for a cheap DVM, or maybe pick up a nice piece of equipment second hand. All seems to be fine, but at the back of your mind you can't help feel that 'seems' is not really the kind of measurement certainty you need. When the meter displays 0.254V how do you know it's telling the truth or just pointlessly impressing you with spurious digits that might have nothing to do with the actual voltage under test.

The answer to this problem is calibration, and this month we have just the project to keep your meters – old, new, cheap or pricey – on the path of honest, repeatable accuracy. The *Multimeter Checker/Calibrator* is a neat, compact design that can help restore your faith in measured values of AC/DC voltage and current, plus it includes a 0.1%-tolerance resistor to check the ohmmeter function on your meter.

A good DVM really is essential for all aspects of electronics, and knowing it is displaying accurate readings should be at the heart of your approach to all measurements.

Time for a 'scope?

Once you have a well-calibrated DVM the very next piece of test gear on your list is of course the oscilloscope – 'scope for short. The problem is there is a truly bewildering array of choices. To help you navigate through the blizzard of options, this month's *KickStart* from Mike Tooley is a great overview of the types to consider, including some suggestions for cheap but reliable second-hand models. High-end scopes can be eye-wateringly expensive, but the good news is you really don't need to spend a fortune. A perfectly useable scope can be had for £100 or less, so don't let cost put you off investing in this most flexible piece of electronic test gear.

Matt Pulzer
Publisher

AI and robots – what could possibly go wrong?

Max the Magnificent

We are currently making tremendous strides in the field of artificial intelligence. There are also a lot of interesting developments in the world of humanoid robots and cobots (collaborative robots) which work hand-in-hand with humans. Let's hope we all stay friends.

As I pen these words, progress is accelerating dramatically on the artificial intelligence (AI) front. The first person to really contemplate the possibility of AI was the English mathematician and writer Lady Ada Lovelace (daughter of the English romantic poet Lord Byron). In her early 20s, Ada assisted the English polymath Charles Babbage when he started work on his proposed general-purpose mechanical computer, the Analytical Steam Engine, in the late 1830s.

Babbage only viewed his Analytical Engine in the context of performing mathematical calculations. In her notes, Ada discussed how the numbers being processed could represent abstract symbols, such as musical notes, and that future versions of the engine 'might compose elaborate and scientific pieces of music of any degree of complexity or extent.'

During WWII, the English mathematician and computer scientist Alan Turing started to ponder the possibilities of machine intelligence. He gave a lecture in 1947 that discussed how machines could learn from experience, and in 1948 he wrote a paper entitled *Intelligent Machines*, which introduced many of the concepts that are central to today's AI. Unfortunately, he failed to publish this paper, which meant most of his ideas had to be reinvented by others later.

The AI ball starts to roll

In 1956, the *Dartmouth Summer Research Project on Artificial Intelligence* was held at Dartmouth College in New Hampshire, US. This seven-week brain-stem-storming session of mathematicians and scientists is widely considered to be the founding event that set the AI ball rolling – and it hasn't stopped rolling since.

Having said this, the ball did not roll very fast at first. Although expert systems, which use knowledge and rules-based approaches, were formally introduced in 1965, progress was painfully slow, largely because computers of the time were limited in memory and performance.

Work on expert systems picked up pace in the 1970s and they really started to proliferate in the 1980s. By the 1990s, however, a lot of us had a sinking feeling that they weren't living up to their promise. Things were not helped when the marketing weenies hopped on the bandwagon and started to stamp 'Artificial Intelligence Inside' labels on everything, even things that had nothing to do with AI whatsoever (in much the same way we currently see 'Gluten Free' stamped on foods that never had a hint of a sniff of a whiff of gluten in the first place).

Just when we least expected it

To be honest, throughout the 2000s, I'd largely relegated thoughts about AI to the recesses of (what I laughingly call) my mind. I knew work was still ongoing in academic circles, but I really didn't envisage any real-world applications for quite some time.

All this started to change in the 2010 to 2015 timeframe when the combination of more powerful computing engines coupled with new AI architectures based on digital artificial neural networks (ANNs) and new AI algorithms such as convolutional neural networks (CNNs) sprang onto the scene.

How big? How fast?

In 2018, an AI research laboratory called OpenAI published a paper titled *AI and Compute*, which defined two eras of AI computation requirements. During the first era, which started with the Dartmouth Workshop and lasted until 2012, the requirements for AI computational capability doubled approximately every two years, which roughly mapped onto the well-known Moore's law.

A 'perfect storm' occurred in 2012 with the introduction of new AI architectures and algorithms. The result was an inflection point across multiple domains (speech, vision, language, games…) that heralded the second (current) era, in which AI computer 'power' started to double every 3.5 months.

How do we do it?

The first AI systems ran on some of the larger computers available at the time. Of course, processor technologies have improved dramatically. Also, AI algorithms, both big and small, have become more sophisticated and more varied. It's now possible to get a humble microcontroller to perform some simple AI functionality. For example, the first AI app I created ran on an Arduino Nano 33 IoT, see: **https://bit.ly/3pJ99Cz**

Until recently, heavy duty AI applications ran on general-purpose field-programmable gate arrays (FPGAs) or graphics processing units (GPUs). Over the past year or so, however, I've talked to a bunch of companies who are developing special analogue, digital and even optical-based devices capable of performing the billions-upon-billions of computations required to implement high-end AI models at extreme speed while consuming relatively little power.

What about robots?

I'm glad you asked. I've also been talking to several companies that are working on humanoid robots. For example, EVE robots from Halodi Robotics are already operate as nighttime security guards in factories. They are also employed in hospitals and supermarkets.

I'm sure you've heard about ChatGPT, the AI chatbot introduced by OpenAI last November. A lot of people are worried about kids using chatbots like this to do their homework, but I don't think we've fully wrapped our brains around all the potential applications (and problems). For example, I recently read about a non-invasive ChatGPT-based system that can translate activity in the human brain into a continuous stream of text: **https://bit.ly/3MqgFKy**

One of my favorite science fiction books is *Great Sky River* by Gregory Benford. Set tens of thousands of years in the future, humans have spread across the Milky Way. When they approach the galactic center, they butt heads (or whatever) with mechanoid civilizations. The 'Mechs' regard biological lifeforms as an infestation to be eradicated. We never do learn what happened to the Mechs' creators, the first of whom had to be biological in nature. I don't know about you, but I can't help thinking: 'AI and robots – what could possibly go wrong?'

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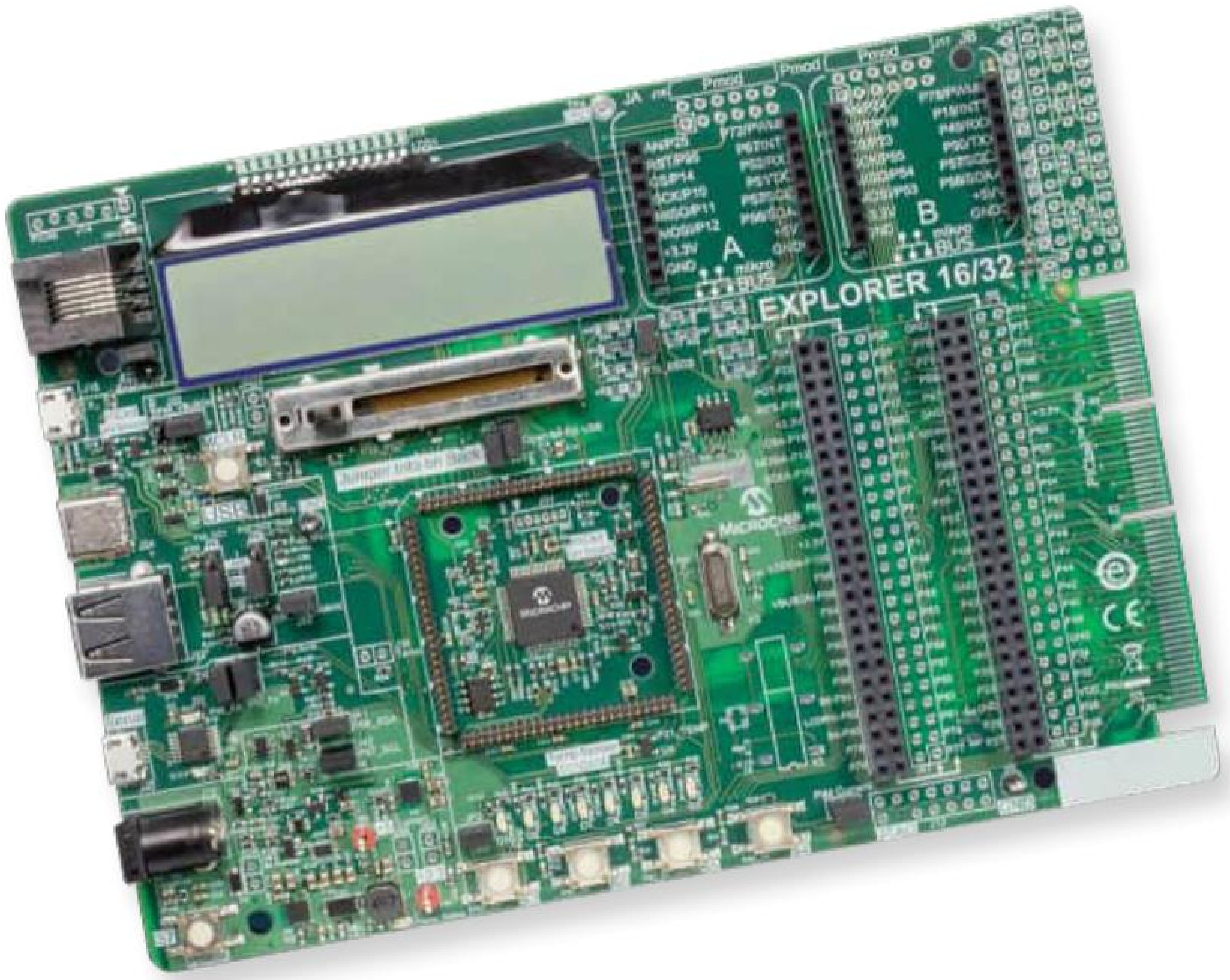

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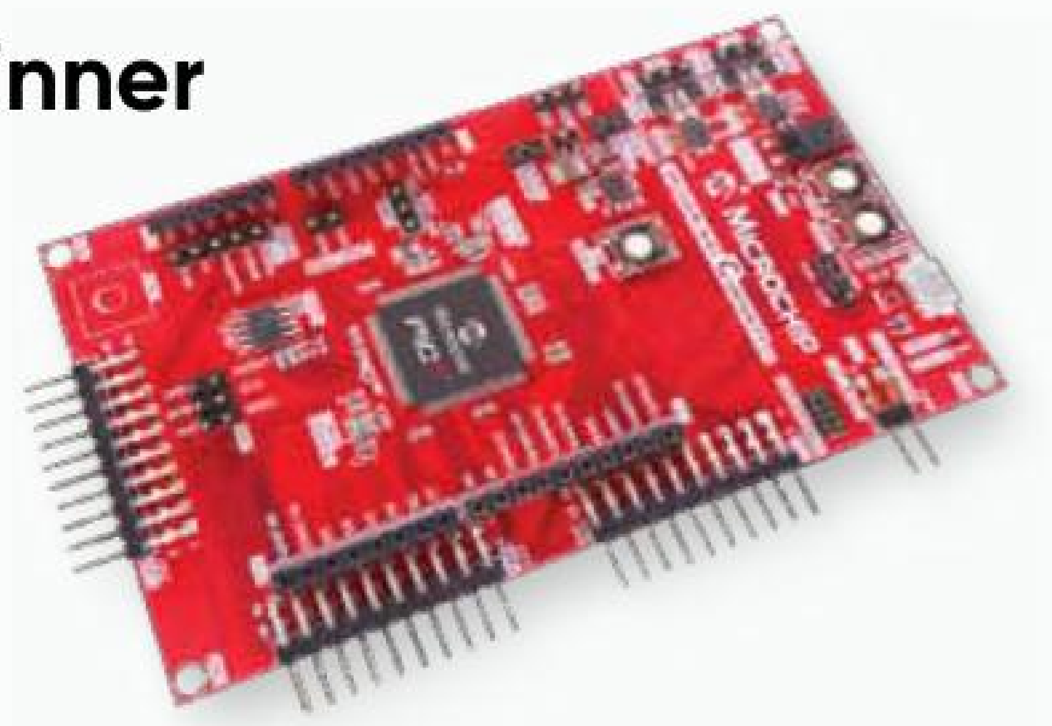

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The **Fox** Report

Barry Fox's technology column

Got a sec?

The International Bureau of Weights and Measures wants to get rid of leap seconds. What are they? And why banish them? It's all about time. The time it takes for the Earth to move round the Sun is not exactly 365 days. So, we have leap years. Every four years February gets an extra day to even things out.

Slowing earth

Likewise, the second was originally defined as a tiny fraction of the time it takes for the Earth to turn once on its axis. But the Earth's rotational speed is not rigidly fixed. The length of the day is increasing by a few milliseconds a century – but why? If you have ever seen 'ice ponds' in a field, you have seen the reason why. The ponds are bowls in the ground caused by the weight of lingering balls of ice. A physicist friend calls it 'post-glacial rebound'. 'Remarkably' he says, '10,000 years after the last Ice Age finished, the Earth's crust is still re-bounding after the removal of the enormous weight of ice. Like an ice skater throwing out her arms, the planet has spun ever more slowly.' All thanks to expansion and the conservation of angular momentum.

Atomic time

To reflect this lack of terrestrial precision in defining the second, in 1967 the second was re-defined by tying it to the vibration frequency of the caesium-133 atom in an atomic clock. In turn, the day has since been re-defined by tying it to the atomically defined second. Because the Earth's rotation time varies unpredictably, our earthly time lords (The International Earth Rotation and Reference Systems Service (IERS) in Paris, France) have, since 1972, been adding a leap second whenever they thought it necessary. This has happened 27 times, on no particular day of the year. This random time warping would crash computers, power grids and satellite navigations systems were it not for the fact that they get their time from publicly accessible time signals, which in turn derive from atomic clocks. Nevertheless, to avoid confusion, leap seconds will be abolished in around ten years' time.

NPL and time

But what would happen if the publicly accessible time signals were suddenly lost or blocked? This is one of the topics now being researched by the UK's NPL (National Physical Laboratory). The NPL occasionally opens its doors for press events and puts on a public relations show, with a few simple science demonstrations and presentations. I went along to one recently, hoping to plug big holes in my understanding of how so much of today's technology relies on precise timekeeping – think GPS location, video time code, audio sampling frequencies and so on.

NPL was founded in 1900 at Bushy House in Teddington by the Royal Society. It is now a commercial company, with a Chief Executive Officer, but is owned by the government Department for Science, Innovation and Technology. It's home to a thousand scientists and engineers, and has the feel of a modern university campus. But, as I found out first-hand, impressive appearances can be misleading. (See the part-pain, part-farce footnote about what happened when I offered NPL's Official Public Relations Coordinator and CIPR-accredited PR practitioner sight of my draft text, before publication, along with a few specific questions. Apart from being ludicrously slow to respond, NPL seemed mainly interested in injecting a dose of civil-service-speak semantics while quietly ripping out passages of likely reader interest – without identifying them and without offering explanation or justification. Only a few points were usefully handled and my technical queries were simply ignored.)

Anyway, back to the NPL, and what do those thousand boffins at NPL actually do? I could broadly sum up NPL's mission by quoting the open-day presenters as seeking, 'measurement – independent and impartial'. However, NPL's PR department now officially prefers to say, 'to provide the measurement capability that underpins the UK's prosperity and quality of life'. OK, so be it. What matters is that current research topics include ionising radiation, 5G, 6G, graphene, underwater acoustics, quantum physics, and time.

Navigation and time

It's easy to forget that longitude measurement (ie, how far east/west you are) depends on knowing the time. Pendulum clocks don't work at sea, which is why John Harrison spent 40 years building the no-pendulum clocks and watch which eventually won him the prize offered by the Longitude Act of 1714. Anyone interested should climb the steep hill in Greenwich Park and visit the observatory's exhibition of clocks. These include all five of Harrison's magnificent time measurement devices. You can also see the Greenwich Meridian, on which is based Greenwich Meantime, which dates from 1880. (And also, Sir Sandford Fleming's global 24-hour zones which start with Greenwich as zero longitude.)

Simple modern clocks and watches rely on the way quartz resonates when an electric pulse is applied. But quartz materials vary, so their time does slightly too. Atoms from any one element, however, are all same, so an atomic clock based on caesium was developed in 1955. Atomic time became the reference standard in 1967, and in 1972 UTC (Co-ordinated Universal Time) replaced GMT. There are now 500 atomic clocks around world, some using hydrogen masers rather than caesium. UTC is disseminated (for instance from a transmitter in Rugby transmitting at 60kHz) and is used to control radio clocks, Internet time and NPL Time, a commercial time service that NPL sells and delivers by fibre to the business and financial sector.

The newest generation of optical atomic clocks are accurate to around one second over the lifetime of the

Universe. This precision means that losing leap seconds is now firmly on the cards, and will remove the last link between UTC and GMT.

GPS vulnerability and time

The GPS systems we all now rely on were originally developed for military use (notably, targeting missiles) but declassified for civilian use. GPS relies on time codes and the loss of a reliable signal can have serious consequences. For example, autonomous vehicles can't function without GPS – missiles miss. Such loss of signal may be due to criminal or military jamming, terrorist attack, solar events (flares), building obstruction or unintentional RF interference.

‘Tetra 2-way radio is worst of all because the GPS signal is weak and ‘in the noise’’ explained a senior product manager at NPL. ‘The US is extremely worried about security. Donald Trump talked about the need for alternatives, and the recent jamming of drones in Ukraine has concentrated minds. The UK economy would lose £1bn over a five-day period if GPS were lost.’

Fibre is of course immune from jamming. But it's vulnerable to workmen digging up the road. And fibre isn't portable. One solution/idea is to use LEO (low earth orbiting) satellites to improve the robustness of time distribution.

‘Resiliency through diversity’ is the NPL approach. ‘The better the time, the better the data. No single solution can provide a fully resilient assured capability’. Arguably one of the most significant, and least talked about, research projects at Teddington is development of a miniature atomic clock which can provide ‘holdover’ – keeping precise time if the GPS signal is lost. Currently, the larger clocks can function with microsecond accuracy for around a month before time drift starts to compromise location accuracy. Fitting drones with atomic clocks could be a military game-changer. Very likely it's already being done, but I suspect NPL regrets talking openly about it.

Mystery time

Sometimes it's not clear where devices are getting their time from. For example, I still use one of the Joggler devices sold well over ten years ago by UK mobile phone operator and Internet service provider O2. Made by OpenPeak and powered by Ubuntu, the Joggler looked like (and worked as) a photo frame but also connected to the Internet and ran some Google Apps and simple games.

Joggler was a commercial flop and most of the online functionality disappeared long ago. But my Joggler still connects to the Internet and works as a clock, getting its time code from some unknown service. The time refreshes after a reboot and correctly observes Summertime shifts. But it's not a radio clock getting time off-air. Somehow, Joggler must still be connecting to some ghost time server, somewhere. NPL either doesn't know, or does not want to say, where devices like old Jogglers are managing to find their time.

NPL Coda

Journalists are often criticised for not letting companies, or individuals, see articles ahead of publication. Some editors expressly forbid it. That's not an attitude I support – fact-checking can help stop rubbish ending up in print or on the Internet. However, it's true that there is a risk that the subject will take so long checking the draft and/or asking for unreasonable changes that deadlines are missed.

Despite this, I thought it would be nice to offer NPL the opportunity to check what I'd written for *PE* for factual accuracy; and while they are at it, answer a couple of queries. I used NPL's press contact form to offer sight of my draft and immediately got back an acknowledgement email assuring, ‘we are working to get back in touch with you as quickly as possible.’

Four days later NPL's PRC (public relations coordinator) and CIPR-accredited PRP (PR practitioner) sprung into action with the re-assurance ‘Yes please! I can get it back to you fairly quickly too.’ I immediately sent the draft, but I heard nothing back so after eight days politely reminded.

‘Apologies for the delay’ said acronym-heavy NPL's PRC and CIPR-APRP, a day later ‘I've asked a colleague in the T&F department at NPL to review your article and answer your questions below. I will get back to you as soon as possible.’

After eight more days I'd heard nothing so politely reminded NPL that some three weeks had clocked up (ironically on the topic of time) ‘since my original contact and I am still waiting for NPL to comment.’

After another four days had elapsed when NPL's PRC and CIPR-APRP offered, ‘Huge apologies for the delay. There has been a lot of colleagues out of the office recently. Let me follow this up now and I will get it back to you early this week.’

Silence and inaction reigned supreme, so after six weeks, or a full lunar month and a half, I spent the now princely price of a First Class Stamp on snail-mailing NPL's CEO Dr Peter Thompson, with the suggestion, ‘If ever you are upset by any writing about NPL that you consider inaccurate, you might like to refer to this timeline.’

A week later an obviously embarrassed NPL staffer phoned to admit that NPL's PR kingpin press contact had gone on leave, without finishing or delegating the job in hand and without even putting an auto-answer message on her email account.

The ‘fact-checked’ draft finally arrived, with a couple of useful minor corrections, some semantic changes, chunks of text excised and my queries completely ignored. Initially, there was none of the customary highlighting used to identify changes in a document. When highlighting was added it somehow managed to miss the censored excisions. At that stage I gave up on NPL's front-of-house press machine.

Silver lining

The upside to the whole time-wasting pantomime is that it will make a very good oven-baked rebuttal the next time anyone pushes that original question – ‘why don't journalists check before publishing?’.

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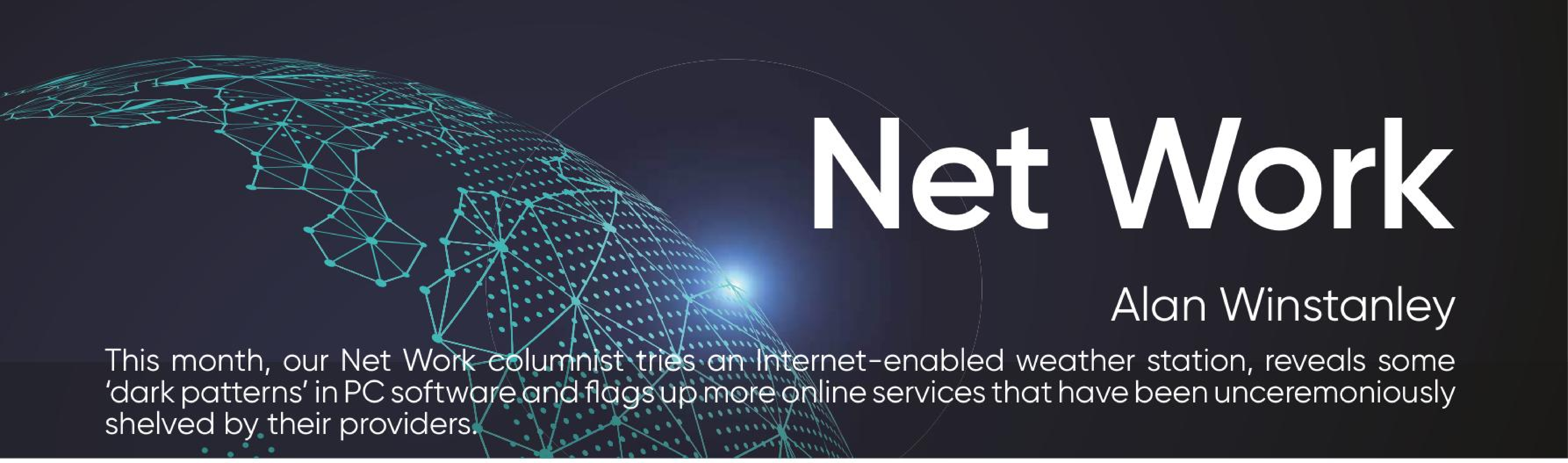

Net Work

Alan Winstanley

This month, our Net Work columnist tries an Internet-enabled weather station, reveals some ‘dark patterns’ in PC software and flags up more online services that have been unceremoniously shelved by their providers.

As the 18th Century saying goes, ‘When two Englishmen meet, their first talk is of the weather’. Our magazine has published a number of projects that appealed to anyone interested in observing weather patterns, such as the December 1995 issue which carried the late John Becker’s *Met Office* (Meteorological Office) weather station. At the time, it was a highly popular design that could upload data to a PC.

Then, in June 1999, came his ingenious *Musical Sundial*. John described it to me as something that played a tune at the top of every hour, triggered when the passing sun cast a shadow onto an opto-electronic ‘sundial’ (a ring of light-dependent resistors). Only John could throw in the term ‘garden gnomon’ – the part of a sundial that casts a shadow onto the dial (nothing to do with garden gnomes, another British obsession). His PIC-powered circuit played the ditty ‘You are my sunshine, my only sunshine’ every hour, weather permitting, and I recall a lot of effort went into transcribing the musical notes into machine code. (Incidentally, source code files for those 1990s+ *EPE* projects are still available from the writer on request.)

Meteorology can be a fascinating hobby in itself, and today a large range of amateur weather stations is available that monitor and log all the common weather parameters. This month’s *Net Work* starts with news of my own trials with a modern weather station to illustrate what’s now typically available, and how the world-wide web can bring comprehensive weather monitoring to everyone.

An Ecowitty answer

There are many identical-looking units on sale, and Chinese brand Ecowitt’s products include weather stations to fit most budgets, some of them being Wi-Fi and Internet-enabled. A number of these are clearly aimed at professional users, but I chose the mid-priced Ecowitt HP2551, which has an outdoor array transmitting data at 868MHz for the UK, over a distance of 100m. It monitors rainfall, wind speed and direction, temperature, humidity, UV index and solar radiation levels. Indoors, a console with a good quality TFT screen displays data, and an indoor sensor checks temperature, humidity and pressure. Data is logged and maximum/minimum data is stored, and alarms can be set. After entering my geographical co-ordinates (found using Google Maps or a compass app), the moon phase was also displayed accurately, along with sunrise and sunset times relative to my corner of England.

The Ecowitt HP2551 is a mid-priced Wi-Fi-enabled weather station with crisp TFT display. Target price of £180 or so.

The system can be enlarged as your interest grows. This model supports up to eight wireless detectors including a wire-probe type of sensor that could check, for example, a fridge freezer or attic. There are optional accessories for some models that can monitor soil temperature (for farming or horticulture) or which float on ponds to alert fishpond keepers of ice formation. A separate ‘leaf wetness’ sensor may help gardeners or greenhouse keepers. Ecowitt’s separate air quality sensor is harder to source, but offers PM2.5 (fine particulate matter <2.5 microns) monitoring and integration into some weather stations. If that aspect is of interest, you can also build your own MOS Air Quality Sensor – see p.36 of this month’s *PE*.

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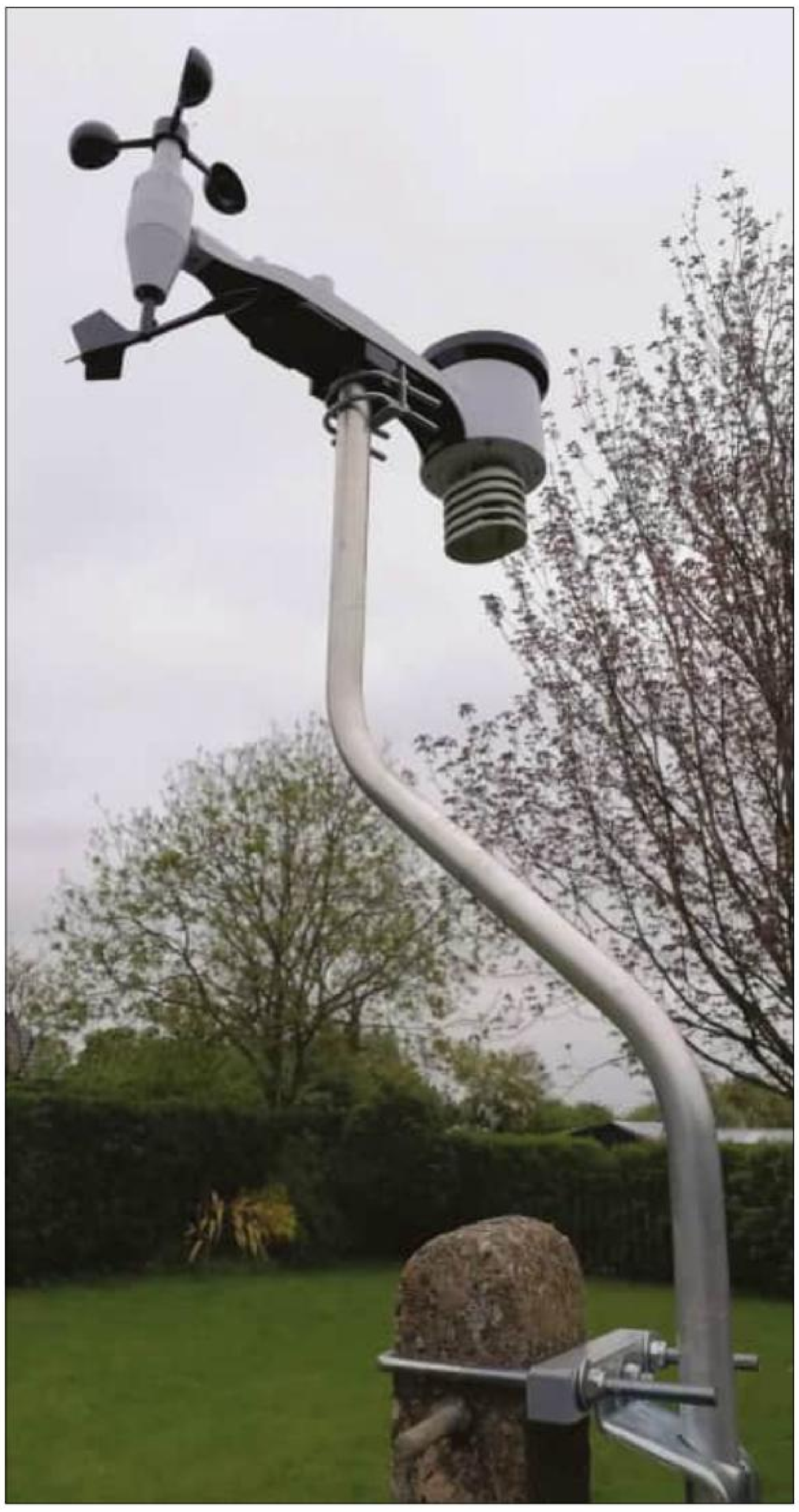
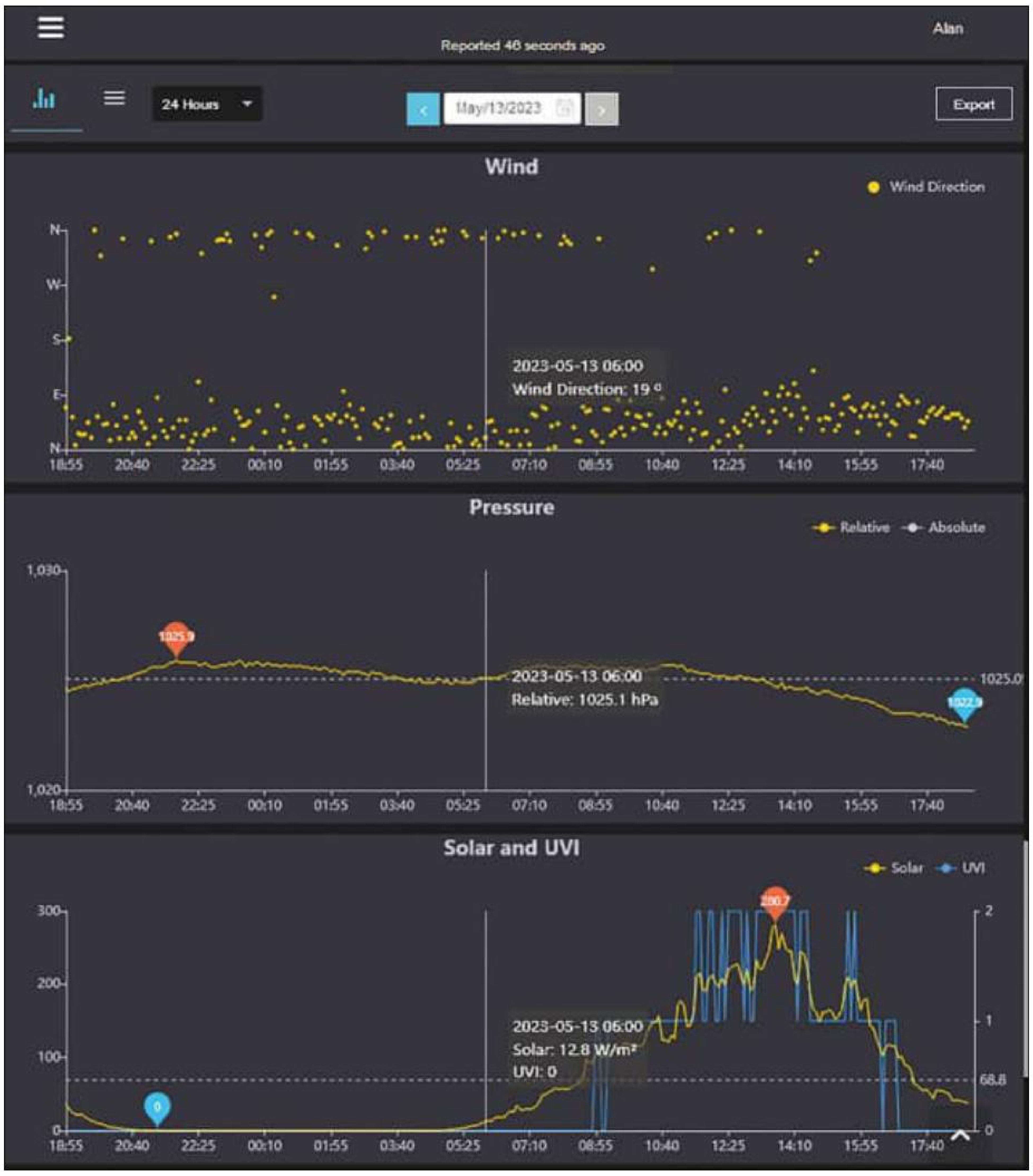


The weather sensor array mounted onto a concrete post in the author's garden. Mast-mounting hardware is available online.



A selection of online graphs drawn from the Ecowitt sensor readings.

Graphic content

I found the high-contrast TFT display to be particularly crisp and clear. Data captured over the past 72 hours can be scoped and displayed in graphs, and it was interesting to see how the wind speed and direction changed over the course of a few days. I could also tell how much rain had fallen overnight, and when. Other factors such as dew point and what wind chill 'feels like' are also available on this model, and some data can be saved to an optional microSD card. I found the instruction manual to be comprehensive and exceptionally clear; it had obviously been written by a knowledgeable English author.

Turning to the practical bits, the display is mains powered and the outdoor array is both solar and battery powered. I use Energizer Ultimate (non-rechargeable) Lithium batteries to power sensors, which are the best available and will last for many years, making them nearly maintenance free. The outdoor array needs fitting to a pole, mast or wall in a reasonably open space. I used an old concrete post to mount a 'cranked' aluminium mast, which I fixed after making some clamping brackets from 20mm aluminium box section and large square U-bolts. The array sat neatly on top, roughly 10 feet off the ground. Importantly, it must point to true North, otherwise the wind vane reading will be inaccurate, so a compass app or Google Maps will help. I discovered a unique range of mounting hardware, U-bolts, clamps, aluminium box section and other useful items sold by ATV in Sheffield, see: **www.aerialsandtv.com** – note that tall aerial masts are not feasible to buy online due to carriage costs, so you might have to improvise or try locally.

With all the hardware finally set up, weather data was soon clocking up on the display and it has proved fault-free so far. Inputting a Wi-Fi SSID means data can be uploaded privately or publicly online, which opens up more possibilities. Trends can then be checked on the Ecowitt app, or a PC browser can log in and view weather data in a web page hosted by Ecowitt, which works really well, though it's less good at forecasting. The Internet connection has been totally reliable to date, and it will optionally upload to Weather Underground and other sites (not tested). I've uploaded more screenshots on the *Net Work Blog* at: **www.electronpublishing.com**

The Ecowitt range is available from Amazon and independent sellers. Amazon's prices can vary wildly and a street price for the HP2551 is about £180. So far, the early signs have been encouraging and the system is proving useful. I've got into the habit of checking to see what the weather's doing before venturing out, or deciding whether I might need a coat. I hope this will encourage readers who are keen on observing weather patterns to set up their own network-enabled weather station: I'm sure John would be nodding furiously with approval!

A dark pattern of behaviour

Last month, I mentioned the design trait of 'dark patterns', techniques used by websites, apps and software that try to manipulate users into navigating down a particular route. Some dark patterns are easily recognised and dismissed, while others are cleverly designed to engineer the user into doing something that they might not ordinarily want to do.

An example surfaced recently when I was rebuilding a Windows PC from scratch. A year or two ago I'd upgraded a legacy Windows 8 PC to Windows 10. All went well until recently, when I decided to transfer it to a new owner. My problem was that Microsoft had tied down the computer's Windows activation key, OneDrive and App store

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2
3
Choose Files
Convert To
Convert Now
Drag & drop files

Zamzar is an invaluable online tool that converts one filetype to another, and is well worth bookmarking.

logins to the current user's Microsoft account, details which I obviously could not share with the new owner.

I carried on preparing the PC regardless but, while installing some software from a USB memory stick, I accidentally nudged the PC's power button which was camouflaged by the black plastic front panel. To my horror the PC shut itself down at a crucial moment, leaving my installation of Corel PaintShop Pro dangling half in, half out: it wouldn't run, nor would it uninstall or re-install, and System Restore made no difference either.

Readers know that sinking feeling as a Windows PC gradually unravels to the point where starting from scratch is the only option left, which is what I had to do. (A tip for next time: hidden deeply in Settings is an advanced power setting option that prevents the Power button from switching off the PC if you press it. I now choose the 'Do nothing' option!)

The challenge was to rebuild the Windows PC while skirting round all the old Microsoft account details. My first port of call was Microsoft's Media Creation Tool, which builds a convenient Windows installer onto a dual-layer DVD or USB memory stick. The Windows 10 installer came from: **https://bit.ly/pe-jul23-ms1**

The final build was 22H2 (ie, 2022, second half of the year). Note that Windows 11 users can instead visit: **https://bit.ly/pe-jul23-ms2**

Back to those dark patterns again. It's common knowledge that, when installing Windows from scratch, the setup routine steers you towards using a Microsoft **outlook.com** account, thereby dedicating that machine and capturing the user, as if there is no other option. That's what happened the first time round, but this time I wanted a clean installation using a so-called 'local account' instead. Deep in a Microsoft forum I found in-depth guidance for a clean re-install, see: **https://tinyurl.com/2cefc55w**

There was just a single line of advice about using local accounts to overcome that nagging Microsoft account problem: during installation and activation, simply choose the option 'I do not have Internet'. So, I didn't connect the PC to Wi-Fi to begin with, just to make sure, and this worked perfectly for me. The PC rebuild went ahead and was problem-free.

Nowhere was it clear that a Microsoft account is not needed during Windows setup, but I successfully configured the new user's details with a Gmail address instead, and no reference to a Microsoft account at all. The PC is now hard at work again with no trace of the previous Microsoft account, achieved simply by disconnecting it from the Internet during setup! On existing systems, you can go to Settings / Accounts / Your Info and choose 'Sign in with a local account instead'.

Getting converted

Another task was to import the new owner's documents into the rebuilt PC. Some files used filetypes that our new software did not recognise. The solution was to use the excellent online converter **zamzar.com** which converted some data into Excel format. Zamzar also converts eBooks, multimedia, CAD, PDFs and many other filetypes online, and is a website well worth bookmarking. Free and paid-for plans are available. Another handy tool is Ninite, which can be configured to fetch and install your favourite choice of apps en masse, see: **https://ninite.com**

Last, on this topic, at about the same time, my Window 10 laptop almost ground to a halt – this is often a sign of large updates downloading and thrashing in the background, and sure enough I was prompted to restart and install updates. As the laptop was Windows 11 compatible, I finally decided to take the plunge and allow it to update to Windows 11. I can report that the update process was completely flawless, albeit slow, with no problems being noted with any existing software running under the refreshed OS. On a Windows PC you can check the status of updates simply by typing 'Updates' after hitting the Win key.

Not so smart car parks

In the last issue, I flagged up the problem of how the demand for cashless payments can discriminate against those who do not, or cannot, use a smartphone. This gulf only ever seems to get wider; one example being the move towards paying with car parking apps on a smartphone as coin-operated ticket machines are withdrawn. This will affect millions of motorists. Customers also need to install the correct app as used by their destination car park. Eventually, a single common app – the National Parking Platform – could be introduced that would work across the country, but I'll believe it when I see it. Early details are at: **https://npp-uk.org**

Vodafone UK recently announced that 3G will gradually be switched off – starting this month – to make way for 4G and 5G services. Other network operators are doing the same, so car park ticket machines and myriad POS (point of sale) devices that use 3G will either have to be upgraded or scrapped anyway. As reported in mainstream media last month, Councillor Nicholas Bennett JP, the Transport portfolio holder at the London Borough of Bromley, said: 'As a pensioner myself, I appreciate that some people have a problem with modern technology. However, we are talking about people who drive a ton and a half of steel, which requires more skill than downloading an app.'

I disagree with this comparison. In many cases, reduced fingertip dexterity or hand-eye co-ordination, working a touchscreen or trying to read the tiny print on a glowing screen means that smartphones are not an option for some people. This does not compare with the skills needed to drive a car. I asked the councillor how non-smartphone owners would use a London car park without coin-operated ticket machines. He replied that residents can register [for car parking access] via the Internet, a home phone or a mobile (cellphone) instead. This only goes to emphasise the relentless march towards a cashless society which will leave behind all those who still depend on cash or can't use smartphones for a variety of reasons. Younger people today, who rely on smartphones, will eventually find themselves in the same boat.

The more things change, the more they stay the same. 1980s fax machines had 'speed dial' buttons that, with just a single button press, sent documents to a stored fax number. This was great until junior staff sent confidential faxes using the wrong speed dial button. There are parallels with today's technology: I've got several namesakes dotted around and occasionally I've received their emails in my own Gmail account. This is because we share similar email addresses and service providers sometimes input contact details wrongly. It does become uncomfortable when I receive details of private dental appointments, long Christmas messages, holiday bookings, building quotations or car insurance policies. These are deleted, of course, but they reveal names, addresses, phone numbers, family details, dentists, insurance brokers and car registration numbers, and I worry a bit about the recipient not receiving important information as well. A young customer assistant at one service provider whom I contacted saw no problem and just shrugged it off. She didn't understand why I was bothering at all, and told me just to, oh, delete everything; a casual attitude to breaches of privacy and data protection that I found quite shocking.

Link account
Open the Google Home app and link your account

The author's Google Home Hub desktop screen was crippled following an update and had to be thrown away.

Android for autos

The author has been grappling with Android Auto, the Google app that mirrors Android smartphone functions onto a car's 'head' unit (the LCD entertainment and info screen). I hoped to try mobile apps such as Waze or Google Maps in-car, although some users say their smartphones get very hot after continued use this way. After many hours of fruitless fiddling, I discovered that, first, my own Honda dashboard no longer connected to Honda's 'Connect' server at all. The digital services supplier whose URL appeared on-screen didn't reply when I quizzed them about the outage, but it turned out that Honda had silently abandoned the Honda Connect service altogether. As the under-developed Honda App centre was mostly useless anyway, I hoped to give the system a new lease of life with Android Auto instead.

It should be a simple case of opening the app and plugging it into the car's USB socket. Things didn't go well there either: although my car's model and year were listed as being compatible with Android Auto, in practice it wasn't, and it was impossible to connect my phone to the car display this way. I learned that my car is now stuck with an Android 4.04 head unit with no chance of running Android Auto on it. As far as online apps go, I'm left with a clunky web browser on the LCD screen (don't worry – it only works when parked) and not much else, as upgrading the head unit or side-loading Android installers is not for the faint hearted. More details of Android Auto for UK users are at: **https://bit.ly/pe-jul23-auto** – noting the compatibility list failed my own checks. It is also strongly recommended that a top-quality USB lead is used. Android Auto is now being built into the latest generation of cars, so using it should not be a problem: just install the app on your phone and connect it to the dash. Apple users can check for Apple Car Play compatibility at: **https://bit.ly/pe-jul23-ios**

Still on the topic of tech makers ditching long-established services that their products (and customers) depend on, owners of the PURE Flow series of DAB/Internet radios (Evoke Flow, Avanti Flow, Sirocco…) which have built-in Wi-Fi and Internet radio connectivity have received the bad news that PURE had suddenly dropped the Flow platform altogether after more than 15 years of service. None of these radios can now receive Internet radio broadcasts as they were designed for the Flow platform only. Many PURE Flow users cherish these otherwise excellent radios, so much so that they go to enormous lengths to replace decade-old fading OLED displays (helped by my own website, which has aided many scores of PURE users to refurbish their radios). PURE is trying to soften the blow by offering £30/€30 discounts off new radios.

Regular *Net Work* readers will recall how the author's Home Hub LCD was 'bricked' following an update, believed to be due to Google's new Fuchsia OS not playing nicely with older kit (unconfirmed). Google was silent on the matter and the perfectly good screen had to be scrapped. Some owners of Google-compatible smart devices are now starting to rue the day they invested in this ecosystem, as one disgruntled user airing his views on Android Authority makes clear. His article certainly makes one think twice about investing in a technology that may be abandoned at a moment's notice, leaving owners with nothing but burnt fingers! See: **https://bit.ly/pe-jul23-dumb**

That's all for this month's *Net Work*. Don't forget that the hyperlinks above are ready-made for you to click on in my *Net Work* blog summary at: **www.electronpublishing.com**

See you next month!

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JEREMY LEACH'S MIDI SYNTHESISER

This advanced MIDI synthesiser is easy to build and can be hooked up to any MIDI-compatible device. It lets you explore the broad range of acoustic elements that capture the characteristics of real and imaginary instruments. It is more than an experiment – it is a full-blown instrument capable of forming rich, detailed sounds using a plethora of settings, envelopes and waveforms – a blank canvas.

The MIDI Spectral Sound Synthesiser uses seven dsPIC chips, each running at 70 or 40MIPS, in combination to produce sounds digitally. This gives it 18 note polyphony (ie, the ability to play 18 different notes simultaneously) and complex sound creation, with 'timbre morphing' being the module's key feature.

It also has low latency, which is important since you don't want an apparent delay between pressing a key on a keyboard and the sound being produced.

Being a standalone sound module, it has the tantalising possibility of being built into custom DIY musical instruments without the need for a computer.

The module is an adventure into real-time sound synthesis, exploring the broad range of acoustic elements that capture the characteristics of real and imaginary instruments. As such, it is a wonderful way of appreciating types of sound, how musical instruments work and why some instruments are notoriously difficult to emulate.

While it is certainly a working device as presented, it's also a great way to experiment with audio synthesis. This is a fun and stimulating pursuit, with an endearing interplay between digital waveform generation and human perception.

The module's design focuses on true parallel processing by splitting the computational load across six 'Tone Processor' chips.

All the source code is available for the firmware and the accompanying Windows desktop software. However, the software is also available pre-built, including any version updates.

This is an advanced project, and there is even a *Technical Reference Manual* for those who want to explore more deeply.

The whole topic of sound synthesis, interwoven with music history, is a rich and intensely interesting evolutionary journey, driven by our instinctive desire to understand and create sound. *The Spectral Sound Synthesiser* taps into that desire.

An overview of the system

Fig.1 shows how the overall system works. It has a MIDI input for receiving MIDI note and control messages (eg, from a keyboard), a USB input for configuration by the Windows software, and a stereo line-level audio output jack, for hooking it up to an amplifier.

You can create patches with the Windows software, and a certain number of these patches are loaded onto the module and stored internally. You can send any tweaks to patch settings immediately to the module and hear the result.

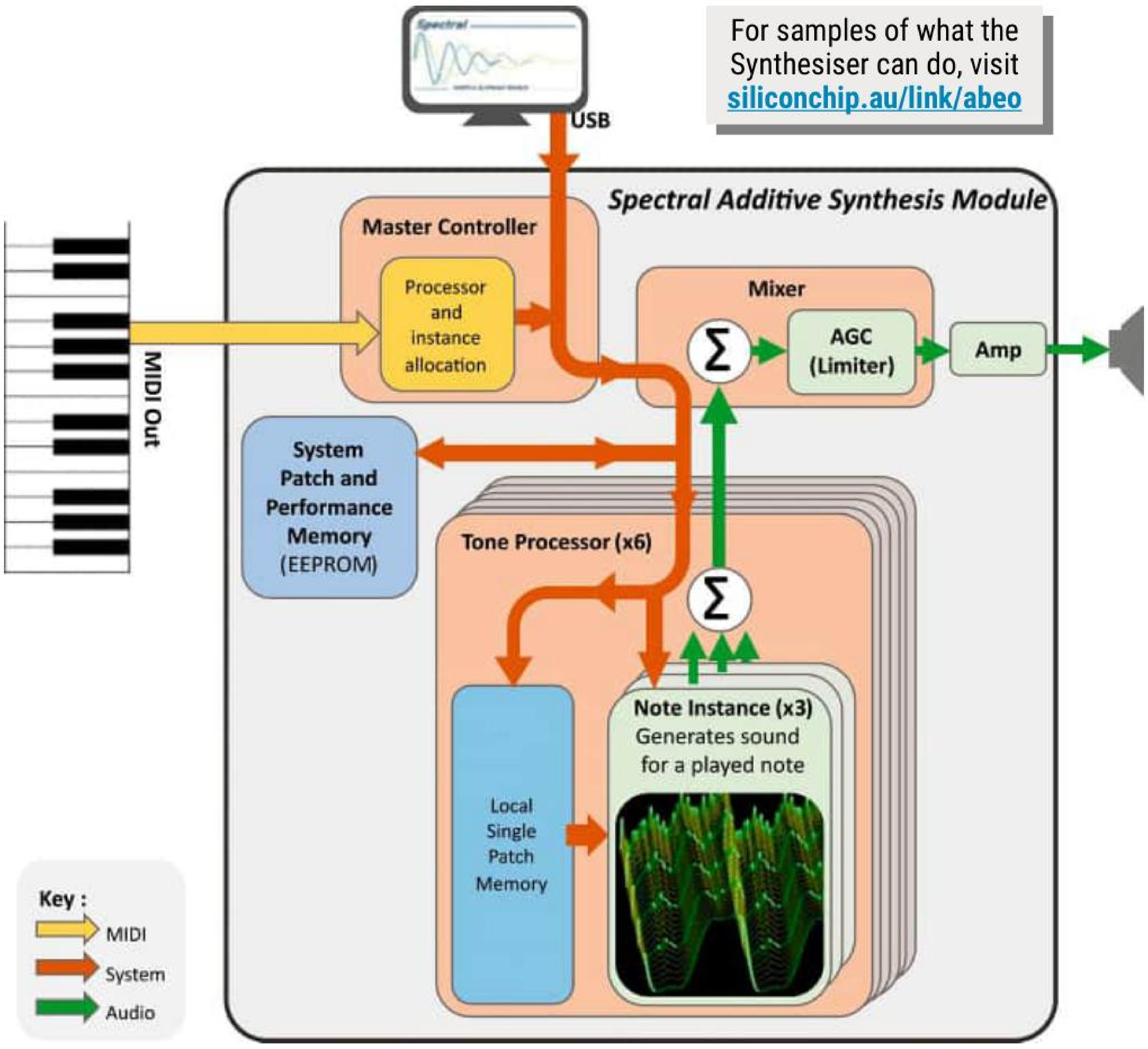
The 'Master Controller' is a PIC18LF25K50 8-bit micro with useful USB connectivity. This chip is common in embedded systems requiring USB. It functions as a hub, processing incoming USB and MIDI messages. It also allocates processors to tasks in the rest of the system.

The six Tone Processors are dsPIC33EP512MC502-I/SP 16-bit chips running identical code to generate digital sound samples. Each calculates up to three live note instances at once, so the system has a maximum of 6 x 3 = 18 note polyphony. Each Tone Processor holds a single patch, but different ICs can have different patches, making this MIDI instrument ‘multi-timbral’.

A single ‘Mixer’ chip, a 16-bit dsPIC33FJ128GP802-I/SP, mixes the samples from all the Tone Processors, limits the generated audio level using automatic gain control (AGC), then passes the audio out through its inbuilt stereo DAC to an MCP6022 op amp.

The output is ‘pseudo stereo’, using a well-known audio trick called the Haas effect, where delaying a copy of a signal from one ear to the other gives a very convincing impression of a stereo field!

The module can hold several patches and ‘performances’ in a 24LC512 EEPROM IC.


USB
Spectral Additive Synthesis Module
Master Controller
Processor and instance allocation
Mixer
Σ
AGC (Limiter)
Amp
MIDI Out
System Patch and Performance Memory (EEPROM)
Tone Processor (x6)
Σ
Note Instance (x3)
Generates sound for a played note
Local Single Patch Memory
Key :
MIDI
System
Audio

For samples of what the Synthesiser can do, visit siliconchip.au/link/abeo

Fig.1: this block diagram shows how the Spectral Sound Synthesiser works. The Master Controller receives MIDI messages from the MIDI In port and patch data from the computer via USB. It commands the six Tone Processors to generate sounds based on the stored patches and possibly stored performance data. These sounds are fed to the Mixer and then to the analogue audio output.

What is additive synthesis?

Ongoing research into hearing and human perception reveals that we are still a long way from completely understanding how our brains process and identify sound. A key element is timbre, which is related to a sound’s frequency spectrum and how it changes over time.

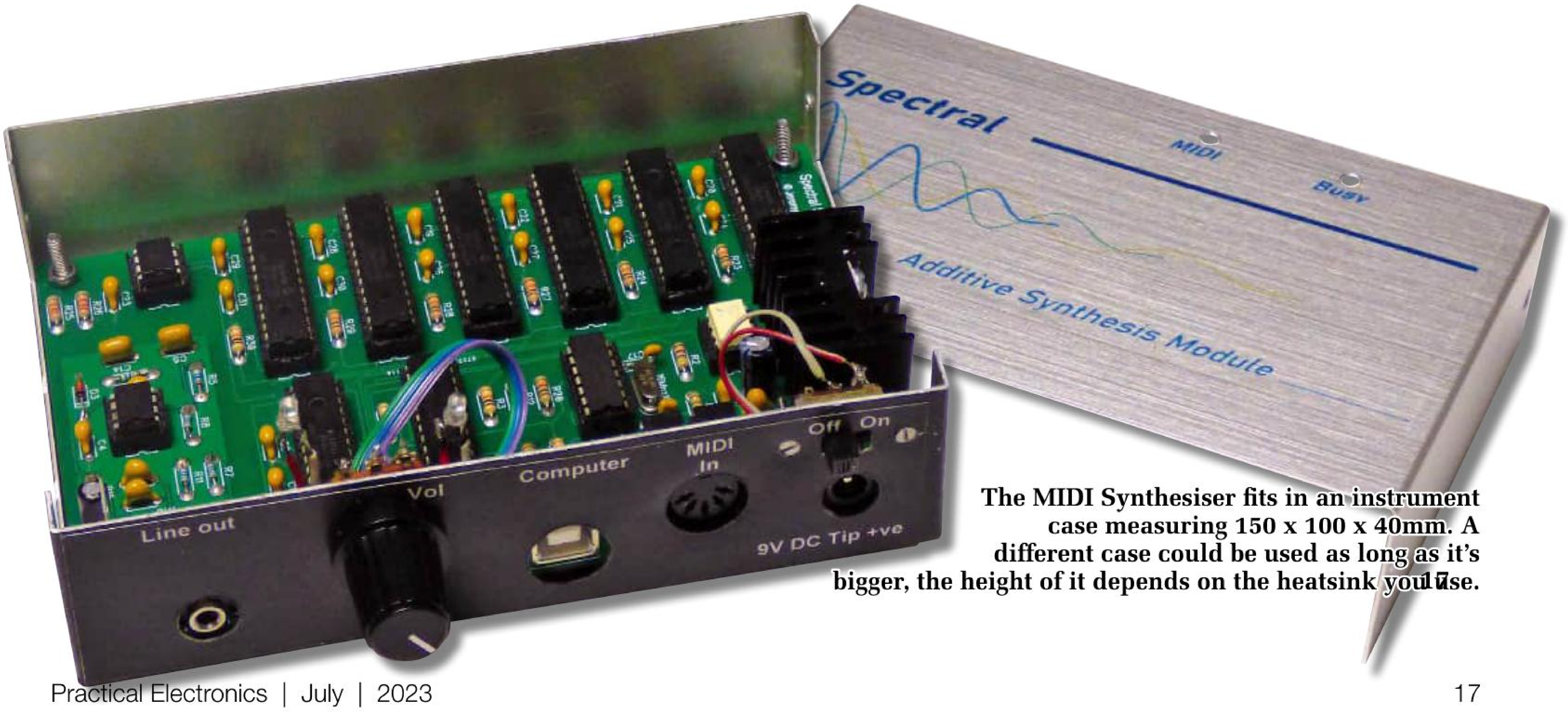
Additive synthesis is a method of creating and modulating timbres based on the fact that any periodic function can be expressed as the sum of a series of sinewaves – the ‘Fourier series’, described by Joseph Fourier in 1822. He was using it to solve heat transfer functions, but this idea soon became widespread, from predicting tides to planetary motion, and much later, audio synthesis.

The simplicity of the idea is appealing because it means that complex timbres can be constructed just by adding sinewaves with appropriate weights and phases.

It turns out that phase is not generally important because our hearing disregards it. This makes sense when pondering sound waves bouncing about in a room; despite the phases of different frequencies getting mangled, we generally do not perceive any timbral difference.

Another appeal is that sounds in nature are based on vibrations where the timbral sinewaves have frequencies that are integer multiples, or harmonics, of the base ‘fundamental’ frequency. This well-defined relationship lends itself to computation. Musical instruments can be recognised by their characteristic harmonic levels, with some examples shown in Fig.2.

The large and evolutionary family tree of electronic synthesisers includes prominent examples of additive synthesis. For example, the well-known and beloved Hammond organ dating back to 1935 stacks tones


Spectral
MIDI
Busy
Additive Synthesis Module
Line out
Vol
Computer
MIDI In
Off On
9V DC Tip +ve

The MIDI Synthesiser fits in an instrument case measuring 150 x 100 x 40mm. A different case could be used as long as it’s bigger, the height of it depends on the heatsink you use.

generated by pickups placed close to rotating mechanical ‘tonewheels’.

Also, the early Fairlight Quasar synth of the 1980s was additive, as were the Synclavier and a few Kawai keyboards. Loom, a modern VST instrument, is also an additive type.

With enough computing power, additive synthesis makes the ‘morphing’ of timbres possible by altering the set of sinewaves being summed over time – akin to what happens all around us with natural sounds.

Additive synthesis also has the great advantage of operating in the frequency domain rather than the time domain. This makes filtering a simple concept, where the filter contour simply scales the levels of the base sinewaves. Brick-wall filtering is nothing more than including or excluding certain sinewaves.

This method of synthesis can create rich, stimulating and captivating sounds. But it has limitations when emulating real instruments compared to sample-based synthesis.

The problem is that natural sound is far more complex than just harmonics; there are ‘in-harmonic’ frequencies in the spectrum, especially for percussive sounds. There is noise from blowing, scraping and scratching. The harmonics are not always exact integer multiples.

So, as a sonic tool, additive synthesis is great. But it cannot always emulate natural sound easily.

The Fairlight CMI synthesiser of the 1980s (which took its name from a Sydney Harbour ferry) was a breakthrough in sound production through sampling. It revolutionised pop music with genuinely new sounds.

The irony is that the inventors started by using additive synthesis, according to co-founder Kim Ryrie (interestingly, also the founder of *ETI* magazine): ‘We regarded using recorded real-life sounds as a compromise – as cheating – and we didn’t feel particularly proud of it.’

This Fairlight model was a ‘sampler’, with the ability to record sound, soon followed by cheaper ‘Romplers’ with recorded sound baked into ROM. These days we can have gigabytes of samples on solid-state hard disks.

It is undeniable that sample-based synths can give amazing results, especially with many nuanced ‘layers’ for parameters such as note velocity. However, they use masses of memory instead of modelling anything on a physical basis. Nevertheless, from the early sampled tape loops of the Mellotron onwards, as used on classics such as the pipe organ in the Beatles’ *Strawberry fields*, it is clear that samples are here to stay.

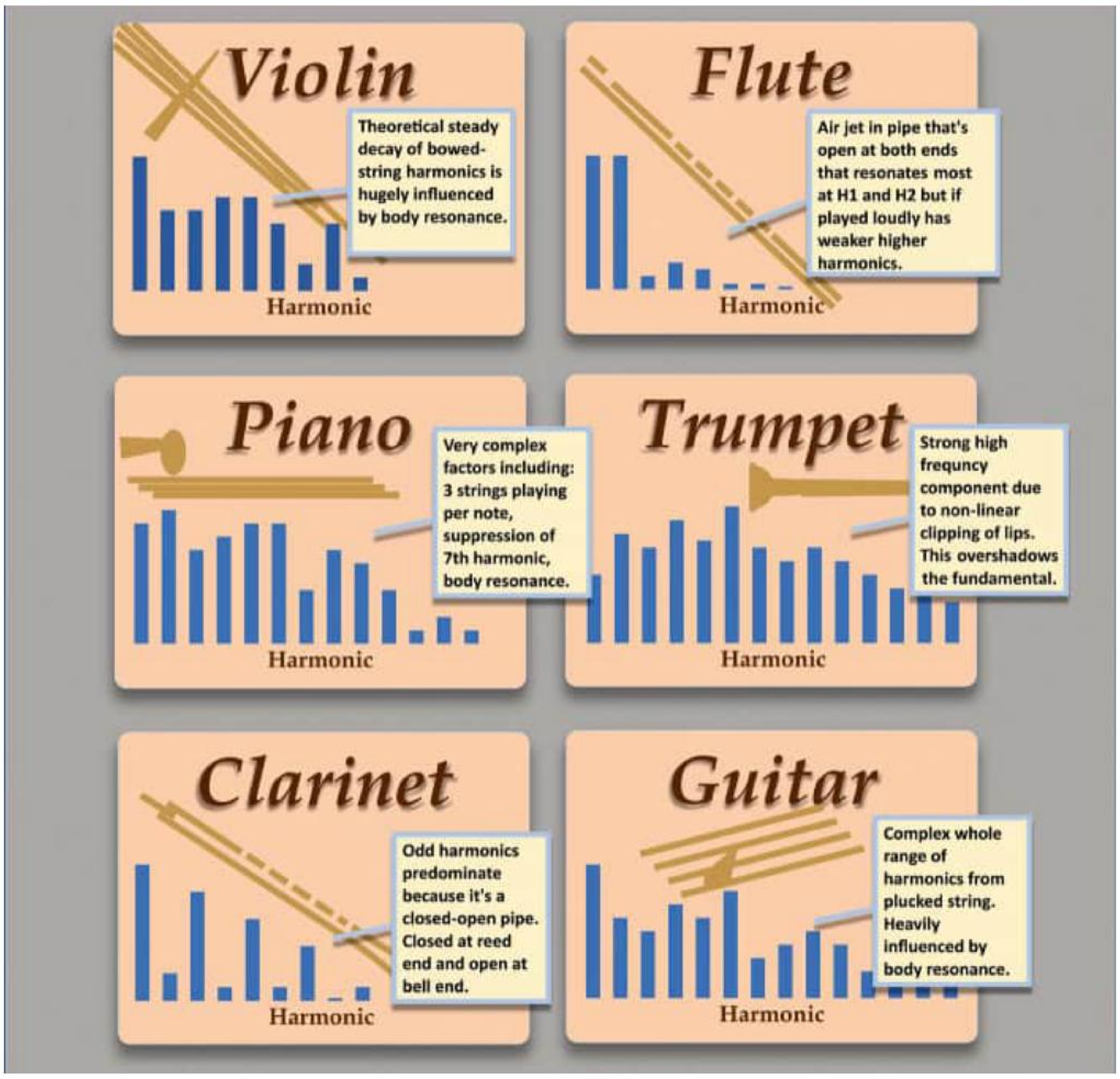


Fig.2: approximations of the harmonic structure of different instruments. From left to right, the bars represent the sequential harmonics above the fundamental. The harmonic structure is what defines the timbre of an instrument, while the fundamental frequency is determined by the pitch of the note being played.

As computing power has steadily increased in recent years, we have seen growth in physically modelled sound, such as in the popular ‘Pianotech’ VST pianos based on the physics of real instruments, ancient and modern. Additive synthesis can also be categorised as physical modelling to a degree because of its timbre-based approach and dynamic nature.

Harmonics and the equal-tempered scale

Real instrument sounds are generated through vibration, such as the movement of air in a flute, the vibration of a guitar string or the oscillating of the skin of a drum. The vibrations create standing waves, having fixed nodes and moving antinodes. The nodes divide the length into equal divisions, leading to the integral harmonics seen in the frequency spectrum of many instruments.

Fig.2 broadly shows how these harmonics have characteristic levels in different instruments, although it is extremely generalised.

An instrument plays at a pitch we recognise as the fundamental frequency, but the tone has a ‘colour’ dictated by the relative strength of the harmonics. The fundamental is known as the first harmonic; the second harmonic is at double the fundamental frequency (an octave higher), the third harmonic at three times the fundamental frequency and so on.

But it is seldom realised that some harmonic frequencies only roughly match the pitches that we recognise in the chromatic (12-note) musical scale! Our brains have heard the pitches of notes from our earliest memories. Yet, the musical scale we use today is relatively recent, and human beings have tried several alternatives, going right back to Pythagoras.

The scale we use today is called the ‘equal-tempered’ scale, where ‘equal’ refers to a fixed frequency ratio between any note and its neighbour. To calculate the frequency of a note a semitone higher, we can multiply by this fixed factor. Since notes one octave apart have a ratio of 2:1, if each semitone has a fixed geometric ratio, that ratio must be the 12th root of two (approximately 1.0595:1).

Remember that this is a human invention to get a system of equal ratios so that music can be transposed without altering how it sounds. Although the oddities of

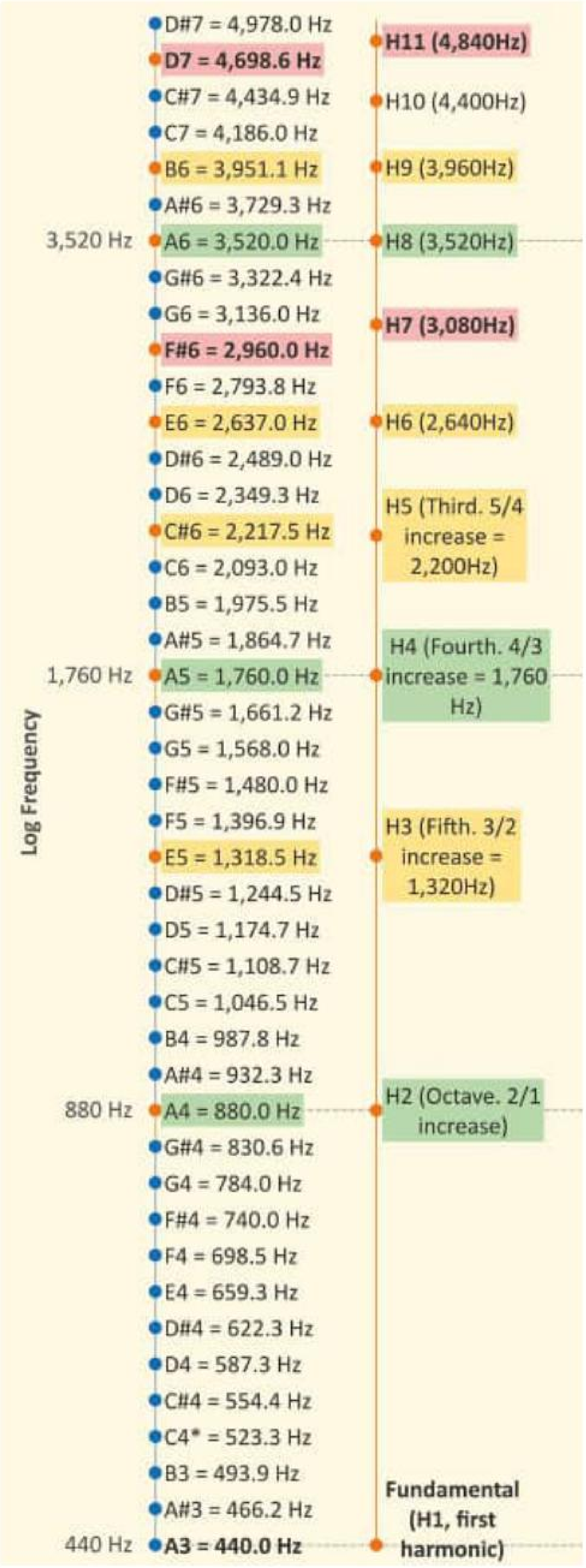


Fig.3: the equal-tempered scale has the advantage that music can be played in any key without retuning the instrument. But some note harmonics do not precisely match any note in the scale, with the worst being the 7th and 11th harmonics. Usually, though, such high harmonics are not especially loud, so this tends not to matter.

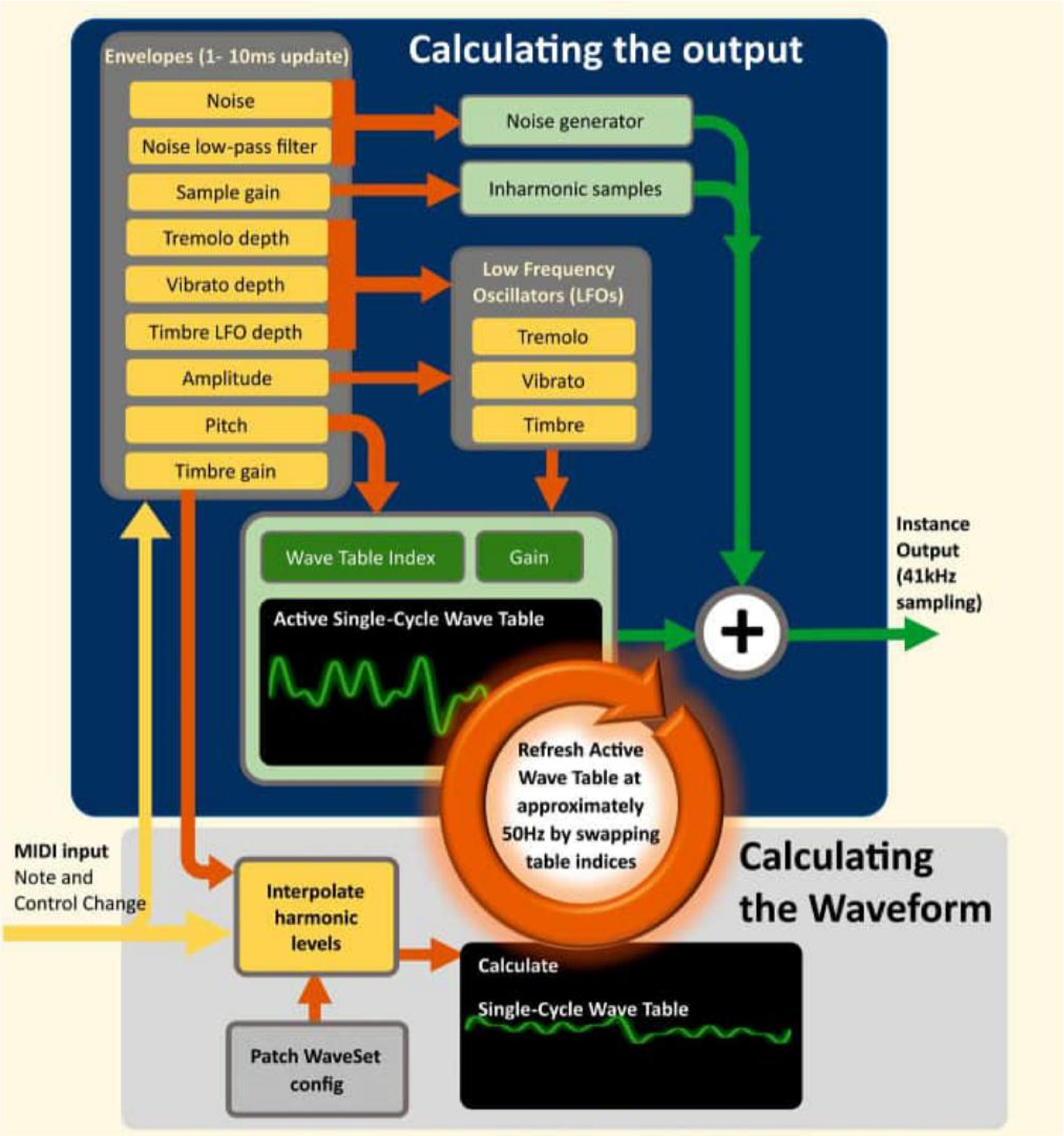

Calculating the output
Envelopes (1- 10ms update)
Noise
Noise low-pass filter
Sample gain
Tremolo depth
Vibrato depth
Timbre LFO depth
Amplitude
Pitch
Timbre gain
Noise generator
Inharmonic samples
Low Frequency Oscillators (LFOs)
Tremolo
Vibrato
Timbre
Wave Table Index
Gain
Active Single-Cycle Wave Table
+
Instance Output (41kHz sampling)
Refresh Active Wave Table at approximately 50Hz by swapping table indices
MIDI input Note and Control Change
Interpolate harmonic levels
Calculating the Waveform
Calculate Single-Cycle Wave Table
Patch WaveSet config

Fig.4: the main tasks and calculations that are constantly being processed by the six Tone Processor chips that do most of the synthesis work.

previous scales contributed to the richness of music diversity, and some bemoan their demise, the equal-tempered scale makes a certain amount of sense.

Fig.3 is a detailed analysis of musical note A3 (440Hz), showing how the harmonics of this note do not always accurately align to the pitches of the equal-tempered scale. Power-of-two harmonics have an exact match in the scale, but others don't (although they are often quite close).

This reveals a degree of 'inharmonicity' in harmonics (ironic, given the name). The analysis applies to all notes; the 7th and 11th harmonics deviate the most from recognisable pitches, and if audible, they sound 'flat' and dissonant.

These imperfections are well-known to instrument makers. For example, pianos are designed to have the hammer strike the strings at the seventh vibrational node to suppress this 'ugly' seventh harmonic! The 11th harmonic is less noticeable, often being naturally quieter.

Tone processors

Fig.4 shows the heart of one of our Tone Processor 'Note Instances', showing how it generates sound. A note instance represents a note we play on the MIDI-connected keyboard. The Master Controller ensures that the played notes are evenly spread across the available Tone Processors.

When a note instance is started on a Tone Processor chip, the first thing that happens is the calculation of the waveform to be used based on the 'static' patch settings, plus other 'dynamic' factors such as the note velocity. This involves looping through all active harmonics and adding corresponding sinewaves together.

Because harmonics are exactly integer multiples of the fundamental, the wavetable holds exactly one cycle of the summed periodic wave. Once that has been calculated, this physical table in memory becomes 'active' for the note instance. To do this, table pointers are swapped so that we never have to copy data slowly and inefficiently from one table to another.

The wavetable for a note instance gets regularly refreshed during its active life cycle. The rate of refresh is not fixed but is approximately 50Hz. As a side note, it is fascinating to read research where they have found that the threshold where humans can detect a change in timbre occurring is often a lower rate than this. Timbre detection is clearly a demanding, abstract recognition task in the brain.

During note generation, several 'gain envelopes' can be applied to aspects of the sound. This is the sound's amplitude envelope. Fig.5 indicates how the system calculates envelopes. Both linear and exponential envelopes can be created, and each section of the ADSR (attack, decay, sustain, release) envelope has a 'target' value. During each section, the envelope's current value moves towards the target.

Exponential curves exist everywhere in nature, relating to energy decay. So, that is a natural choice, and not just for amplitude. For example, when plucking a string, the high-frequency content decays first. The

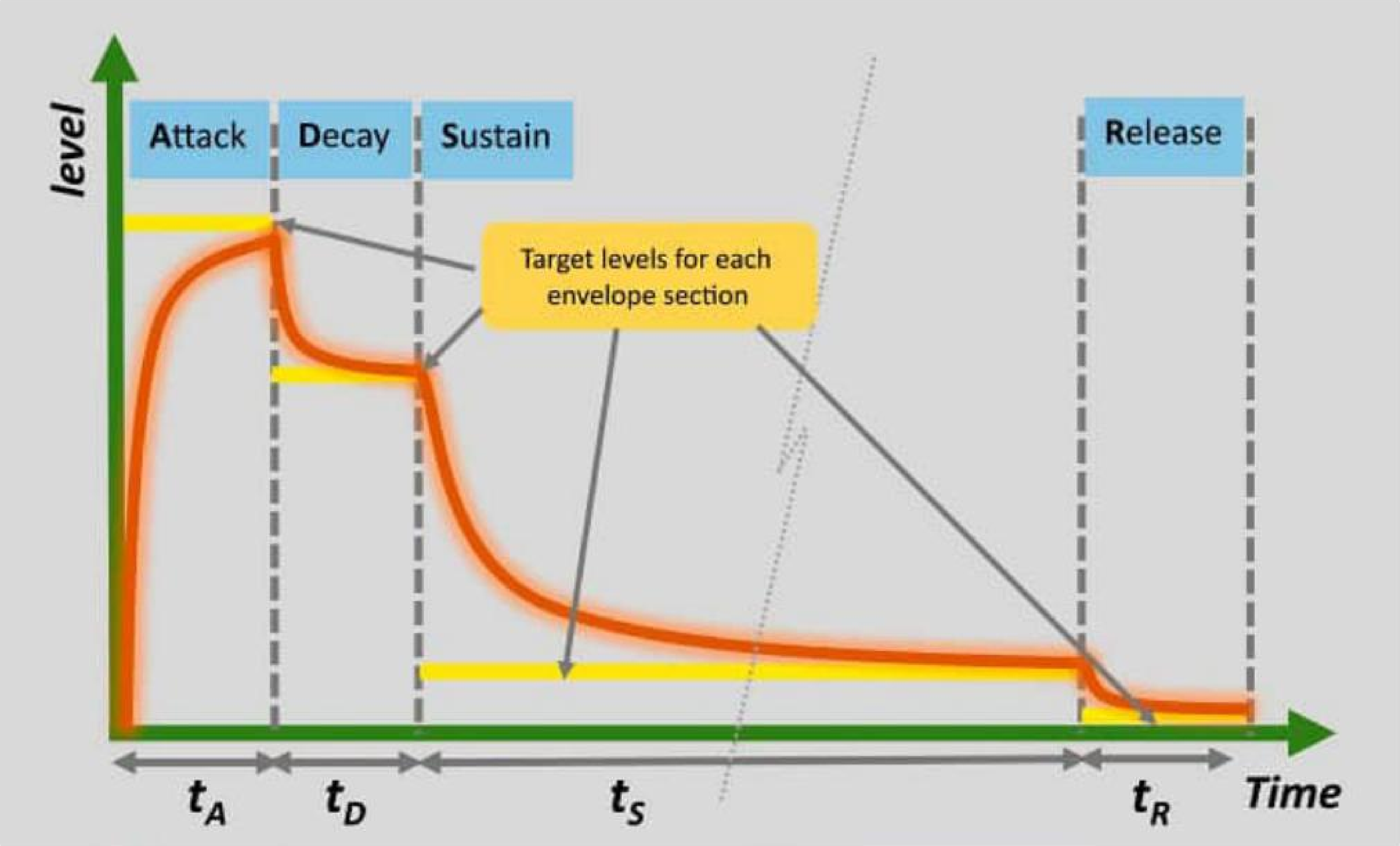


Fig.5: envelopes are time-based profiles that can be applied to various synthesis parameters. The easiest to understand is the volume envelope, which varies the loudness of the note from the time it is triggered until it is no longer audible.

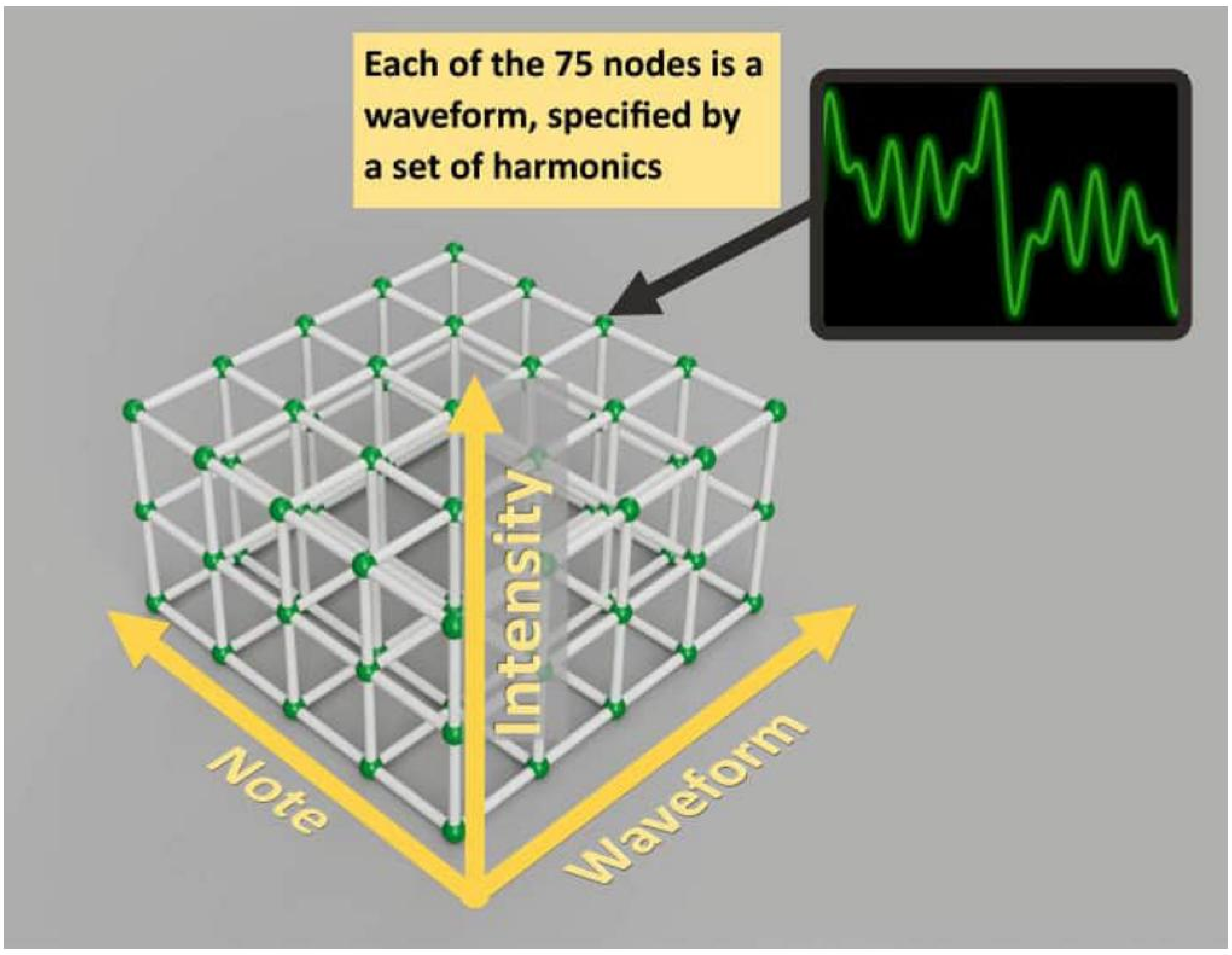

Each of the 75 nodes is a waveform, specified by a set of harmonics
Intensity
Note
Waveform

Fig.6: '3D timbre morphing' is a solution to the problem that the harmonics of various instruments can vary depending on which note is being played, how hard it is being played and so on. This is especially obvious on instruments like pianos, where each key can have a unique sound, and louder notes can trigger various resonances.

jostling of atoms in high-frequency vibration will use up energy at a higher rate.

A note instance also features three 'Low-Frequency Oscillators' or LFOs: Vibrato varies the pitch, Tremolo the amplitude and Timbre the harmonic levels. LFO modulation can add so much character to sound. There are also envelopes for the depth of this modulation, allowing, for example, a gradual onset of vibrato.

Another interesting point is that whereas many synths would use Tremolo across the total sound output of the synth, this module modulates per note, making the overall sound more complex and interesting.

Timbre morphing

Determining the harmonic levels to use when constructing a waveform is quite a complicated process. Fig.6 shows that a patch holds the harmonic data for 75 waveforms, with the waveform to synthesise depending on three parameters. The 'Note' parameter is the position of the note on the keyboard. The 'Intensity' parameter most often means note velocity. The 'Waveform' picks from five waveforms.

Each point in this conceptual 3D space grid has a set of harmonic levels defining a waveform. The current parameter values define the required point internal to this space, and the harmonic levels to use are interpolated from the nearest defined grid points in this space.

Taking this further by modulating through this space with the Timbre LFO can give impressive realism compared to plain vibrato. Typical vibrato purely modulates the pitch of the whole waveform, whereas timbre modulation is harmonic-based and therefore, a far more complex modulation for our brains to perceive.

It is thrilling to hear this difference and realise that our brains feed off the interest in sound. Perhaps, considering the incredibly clever processing that our brains can perform with language, pattern detection and all aspects of sound, this realisation should not be too surprising.

Towards greater realism

We have already mentioned that natural sound includes in-harmonic elements, which do not fit the neat integer multiples of harmonic frequencies. In an attempt to address this point, this synthesiser includes some additional features outside of the purely additive-synthesis approach.

First, a noise envelope is available to help simulate 'blown' instruments. This is a white noise generator with an adjustable low-pass filter.

Second, you can add short in-harmonic samples. These are hard-coded clips of the sound of taps, scratches, clicks and bonks. However, the sample feature also includes an implementation of the well-known 'Karplus Strong' delay line technique of plucked string synthesis.

The 1983 paper by Kevin Karplus and Alex Strong entitled 'Digital Synthesis of Plucked-String and Drum Timbres' first described this technique. It is a computationally simple but effective method of generating realistic, decaying string sounds that start off life in the delay buffer as noise. It adds a powerful tool to this module, even though it does not have anything to do with additive synthesis!

There are also settings to randomly or systematically detune the frequency of played notes, in an attempt to introduce the impurities of real instruments. Plus, there is an option of using two wavetable oscillators per note instance, detuned by a specified amount, giving a chorus-like effect, which tries to account for the fact that instruments like the piano use multiple detuned strings per note.

A final feature is the 'Body Resonance Filter'. This attempts to emulate the body resonance of a real instrument by filtering the overall system sound. After specifying the filter contour in the app, it scales the harmonic levels. Although this method has great theoretical appeal, it has mathematical limitations.

Despite all these extra features, there are still real-life complexities that the module just cannot tackle. For example, a piano has peculiarities due to the stiffness of its strings, where the higher harmonics get sharper compared to the expected harmonics of the string.

This is because the stiffness effectively shortens the string for higher frequencies, raising the harmonic oscillation frequency. This effect applies to all stringed instruments, and is something our system cannot address because the model relies entirely on integral harmonics.

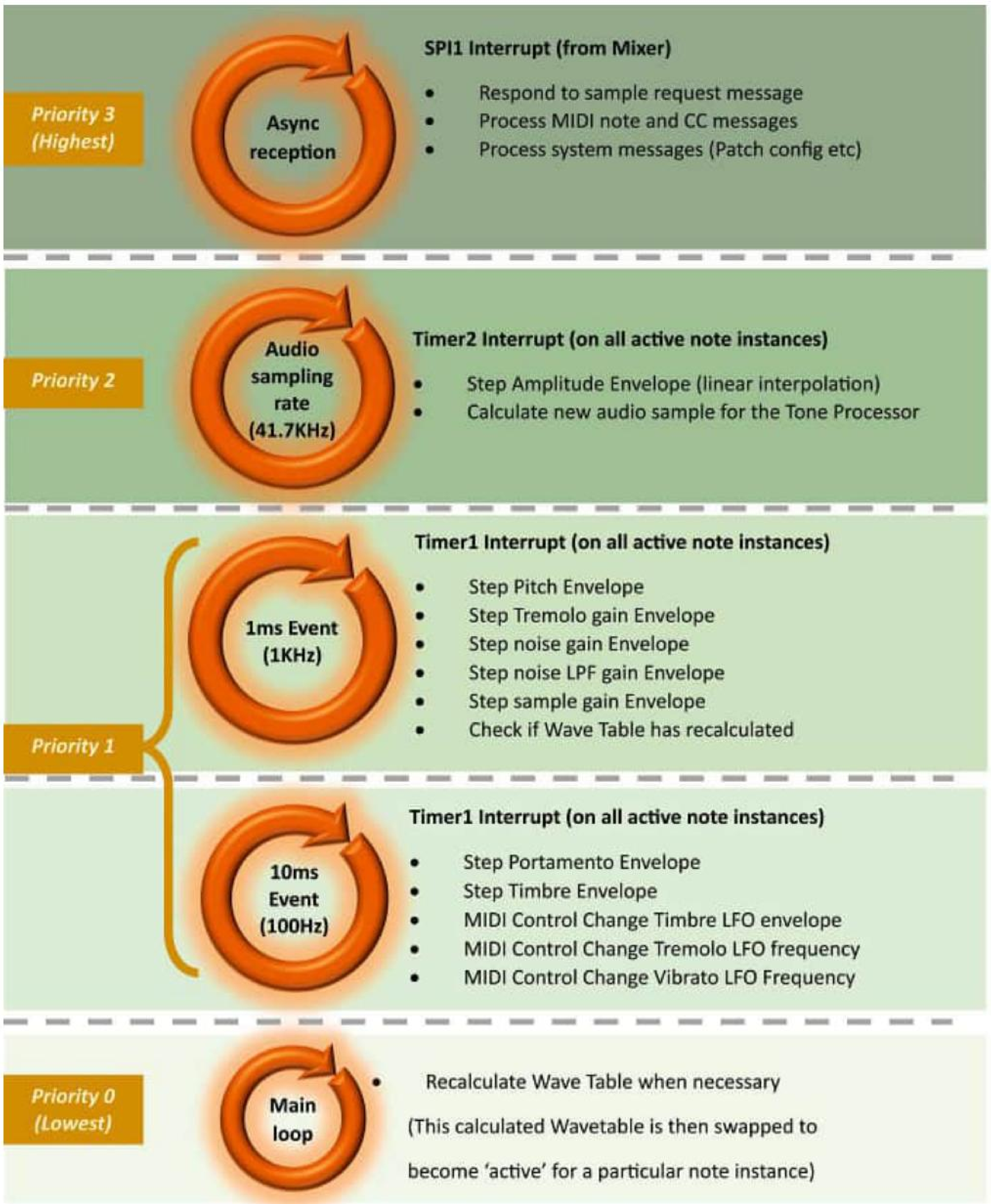

Priority 3 (Highest)
Async reception
SPI1 Interrupt (from Mixer)
Respond to sample request message
Process MIDI note and CC messages
Process system messages (Patch config etc)
Priority 2
Audio sampling rate (41.7KHz)
Timer2 Interrupt (on all active note instances)
Step Amplitude Envelope (linear interpolation)
Calculate new audio sample for the Tone Processor
Priority 1
1ms Event (1KHz)
Timer1 Interrupt (on all active note instances)
Step Pitch Envelope
Step Tremolo gain Envelope
Step noise gain Envelope
Step noise LPF gain Envelope
Step sample gain Envelope
Check if Wave Table has recalculated
10ms Event (100Hz)
Timer1 Interrupt (on all active note instances)
Step Portamento Envelope
Step Timbre Envelope
MIDI Control Change Timbre LFO envelope
MIDI Control Change Tremolo LFO frequency
MIDI Control Change Vibrato LFO Frequency
Priority 0 (Lowest)
Main loop
Recalculate Wave Table when necessary
(This calculated Wavetable is then swapped to become 'active' for a particular note instance)

Fig.7: an overview of all the tasks that the Tone Processor chip has to handle, in order of priority. The highest priority events are those that would cause the sound to break up or otherwise give unexpected results if delayed.

A real-time system

The whole of the module is an example of a 'hard' real-time system, where the deadlines of sample production and processing are immovable. This presents considerable challenges, and the development of the module was a slow evolution of coding, measuring, refining and sometimes redesigning.

The Tone Processors all run entirely in parallel and are polled by the mixer chip to provide samples at the audio sampling rate of 41.7kHz. Inside each Tone Processor is a hierarchy of interrupts, as shown in Fig.7, made possible by the dsPIC's ability to assign priorities.

The main routine of a Tone Processor, the centrepiece of the entire system, is just a simple loop that recalculates wavetables. This 'background' task is unpredictable in duration, is not on a deadline, and can vary depending on the interrupt activity and the complexity of the waveform being built.

This means that the timbre refresh rate could slow down in certain circumstances – although, in use, performance is very acceptable, and timbre changes are perceived as fluid and smooth.

The processing 'layers' above this base main loop are concerned with calculating envelope steps, calculating the sample output and processing received data.

A trick used to improve throughput on the SPI bus between the Tone Processors and the Mixer is only sending the changes in sample values. The summing of sines in a Tone Processor can result in a total value exceeding 16 bits. The total on the Tone Processor is a 32-bit signed integer, but the Tone Processor only sends the change in this total, capped at 16 bits, to the Mixer.

This can cause signal distortion, but statistically, this will happen rarely. The performance advantage of this method massively outweighs rare anomalies that are probably not even noticeable.

The software and hardware are designed for speed. All the dsPICs are running at their fastest. All calculations are integer-based, coupled with extensive use of the on-chip hardware multiplier via the compiler's '__builtin' commands. The code extensively uses shift operations for fast multiplication and division, and numerous lookup tables are used, including a detailed sine lookup.

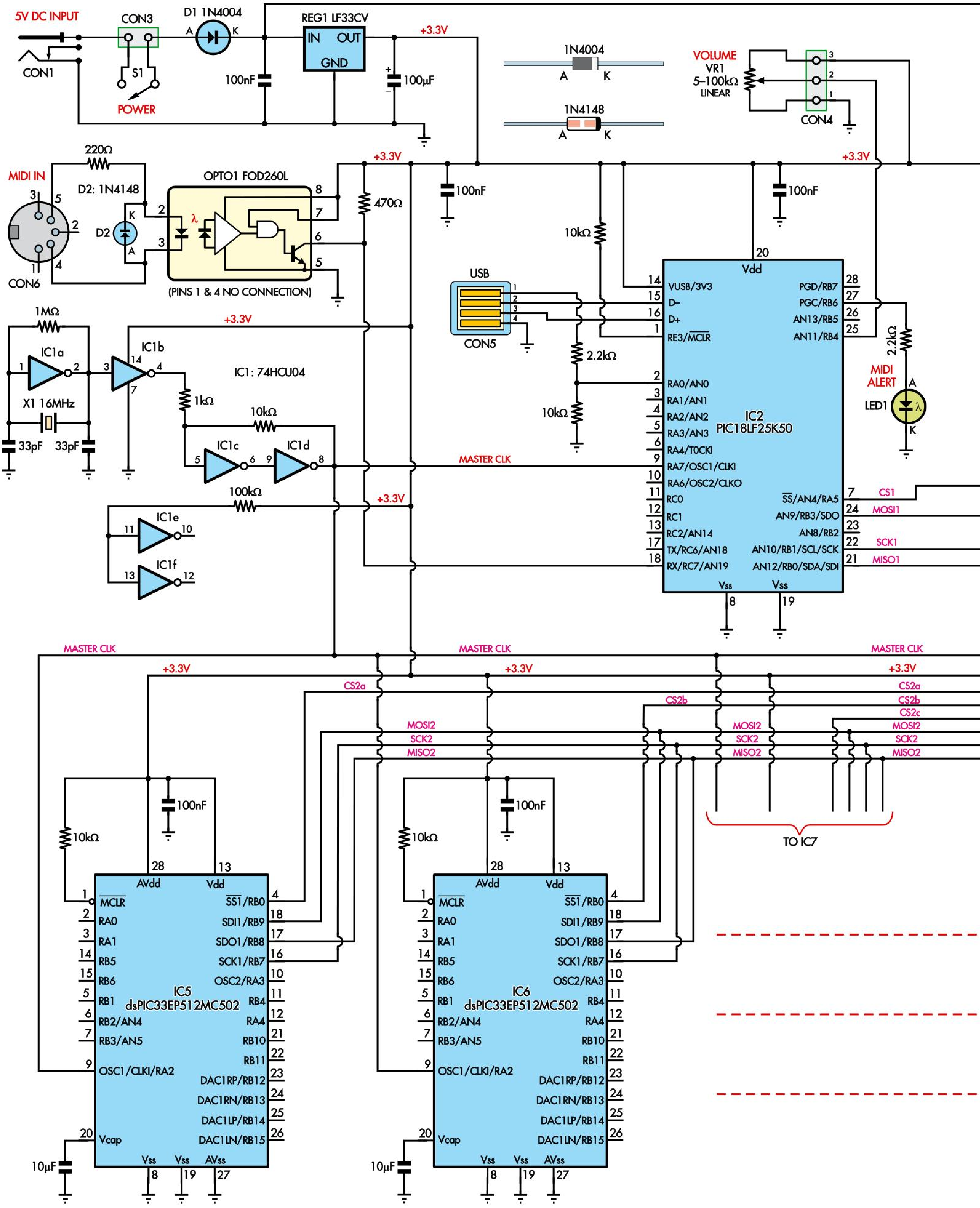
Circuit details

The full circuit is shown in Fig.8. The first thing to note is that the six Tone Processors (IC5-IC10) are identical dsPICs configured in the same fashion, each with just a handful of associated components: one V_{DD} bypass capacitor, one V_{CAP} capacitor (required for the chip's internal regulator) and one 10kΩ $\overline{MCLR}$ pull-up resistor to prevent spurious resets.

Besides the power supply, the only connections to these chips are a common SPI bus, as they are pure number crunchers, and all commands and data are sent on this bus. The only differences between the connections to these chips are that each Tone Processor's CS2a-CS2f input (pin 4) connects to a different select line on the Mixer, IC3.

The Mixer is a different (but related) type of dsPIC processor. Besides being connected to this SPI bus and the six chip select lines for the Tone Processors, it also has two differential analogue outputs from an internal stereo DAC.

These signals are fed to op amp IC4, which converts the differential signals to single-ended audio signals suitable for feeding to the CON2 output jack. Simultaneously, this circuitry filters out the DAC step artefacts using low-pass filters built from added capacitors and the existing gain-setting resistors.


5V DC INPUT
CON1
CON3
S1
POWER
D1 1N4004
A
K
100nF
REG1 LF33CV
IN
OUT
GND
+3.3V
100µF
1N4004
1N4148
VOLUME
VR1
5–100kΩ
LINEAR
CON4
220Ω
MIDI IN
CON6
D2: 1N4148
D2
OPTO1 FOD260L
(PINS 1 & 4 NO CONNECTION)
470Ω
100nF
10kΩ
USB
CON5
2.2kΩ
10kΩ
1MΩ
IC1a
IC1b
IC1: 74HCU04
X1 16MHz
33pF
33pF
1kΩ
10kΩ
IC1c
IC1d
100kΩ
IC1e
IC1f
MASTER CLK
Vdd
VUSB/3V3
D–
D+
RE3/MCLR
RA0/AN0
RA1/AN1
RA2/AN2
RA3/AN3
RA4/T0CKI
RA7/OSC1/CLKI
RA6/OSC2/CLKO
RC0
RC1
RC2/AN14
TX/RC6/AN18
RX/RC7/AN19
IC2
PIC18LF25K50
PGD/RB7
PGC/RB6
AN13/RB5
AN11/RB4
SS/AN4/RA5
AN9/RB3/SDO
AN8/RB2
AN10/RB1/SCL/SCK
AN12/RB0/SDA/SDI
Vss
2.2kΩ
MIDI ALERT
LED1
CS1
MOSI1
SCK1
MISO1
CS2a
CS2b
CS2c
MOSI2
SCK2
MISO2
TO IC7
100nF
10kΩ
AVdd
MCLR
RA0
RA1
RB5
RB6
RB1
RB2/AN4
RB3/AN5
OSC1/CLKI/RA2
Vcap
10µF
SS1/RB0
SDI1/RB9
SDO1/RB8
SCK1/RB7
OSC2/RA3
RB4
RA4
RB10
RB11
DAC1RP/RB12
DAC1RN/RB13
DAC1LP/RB14
DAC1LN/RB15
AVss
IC5
dsPIC33EP512MC502
IC6
dsPIC33EP512MC502

Spectral Sound Synthesiser

A virtual 'half-supply' rail is generated using zener diode ZD1 biased from the +3.3V rail so that the audio signals from IC3 remain within the supply rails of the op amp.

Mixer IC3 also connects to the 24LC512 EEPROM (IC11) using a two-wire I^{2}C serial interface (SDA and SCL). That chip has its own bypass capacitor plus pull-up resistors for those serial lines, and that's it. The last task for IC3 is to drive the Mixer Alert LED, LED2, from its RA0 output (pin 2).

MIDI input, USB and other control tasks fall on the PIC18LF25K50, IC2. It monitors the presence of USB 5V at its RA0 digital input (pin 2) via a 2.2kΩ/10kΩ 'divider', which

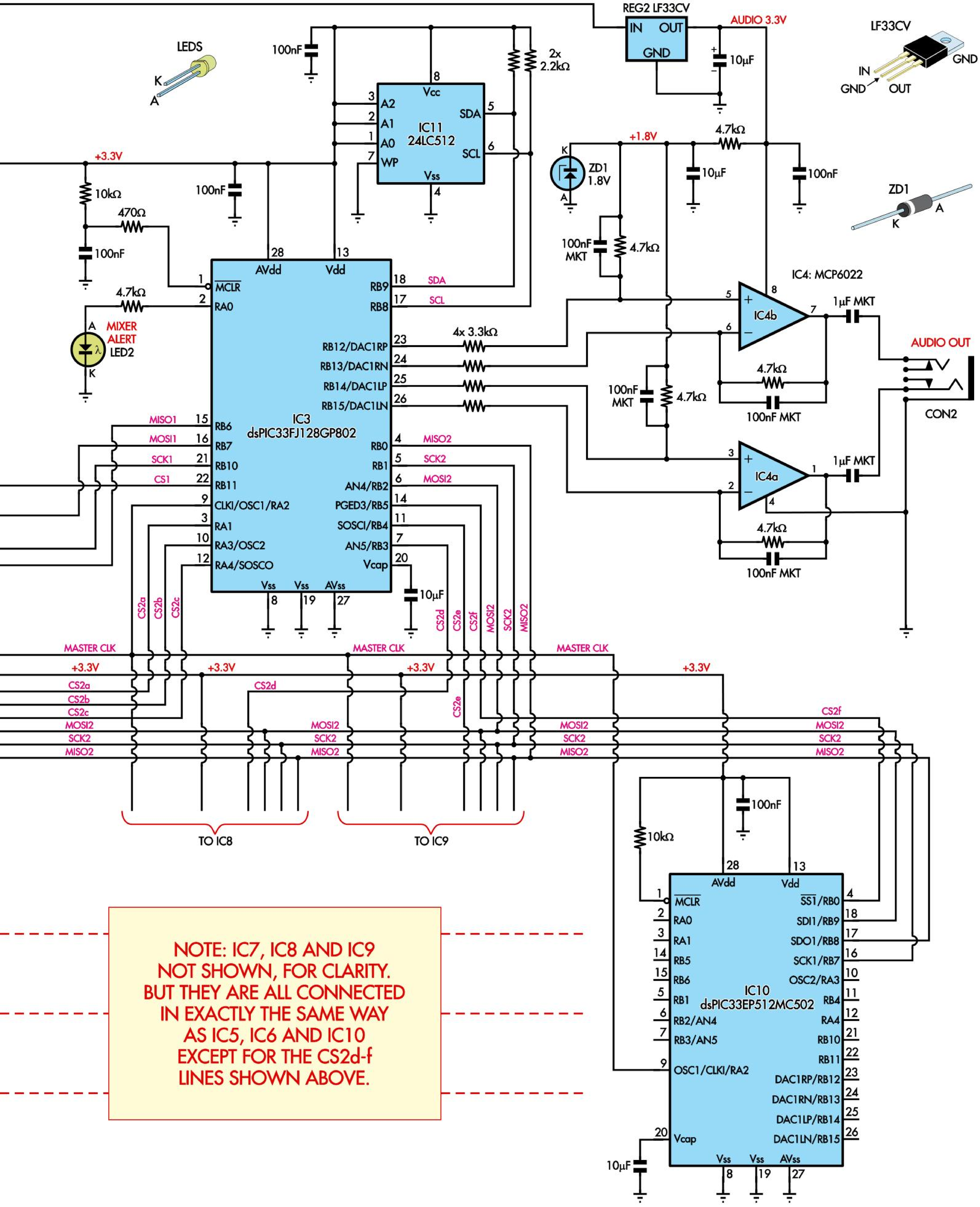
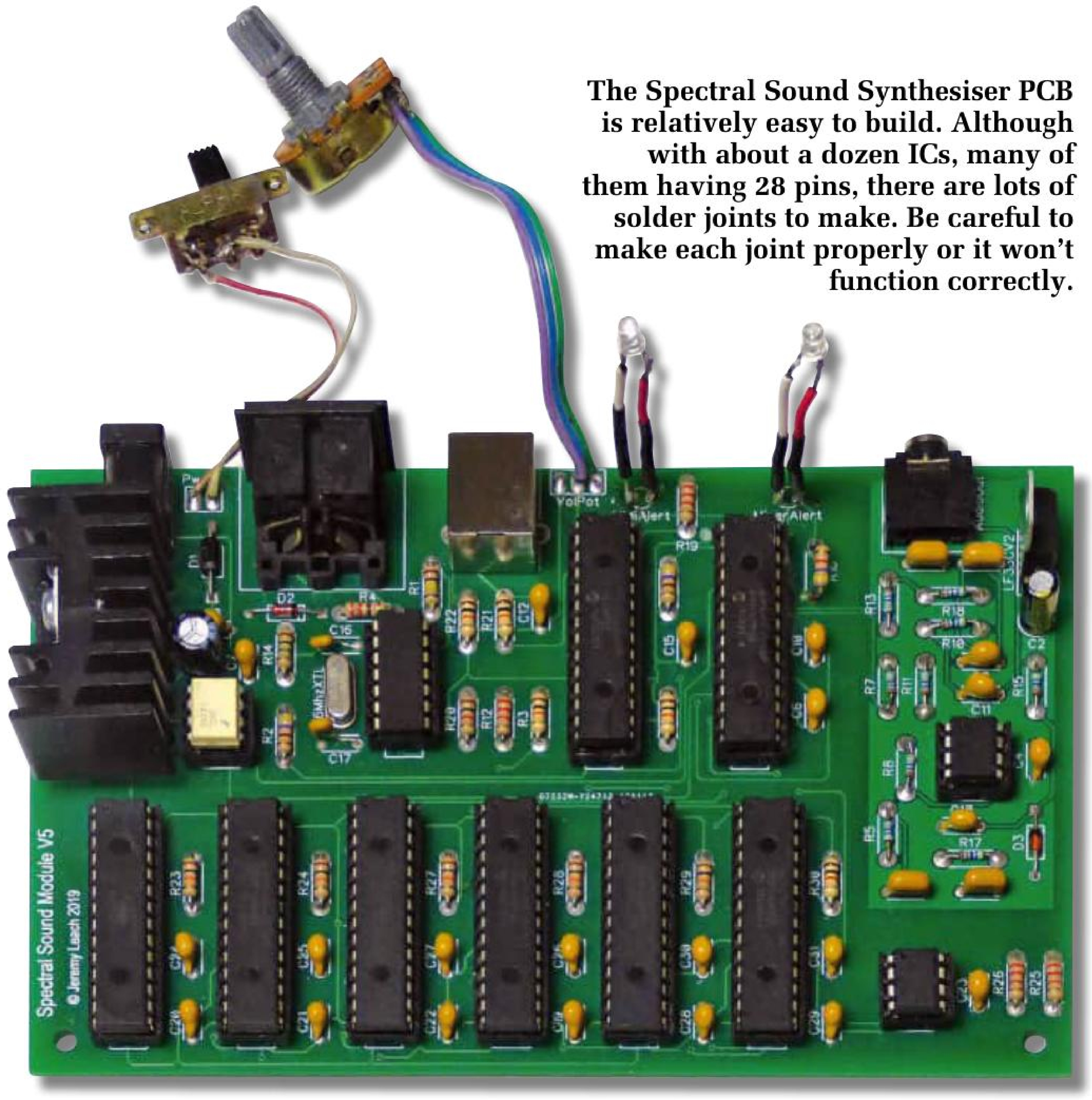

REG2 LF33CV
IN
OUT
GND
AUDIO 3.3V
10µF
LF33CV
IN
GND
OUT
GND
LEDS
K
A
100nF
2x 2.2kΩ
8
Vcc
3 A2
2 A1
1 A0
7 WP
IC11 24LC512
SDA 5
SCL 6
Vss 4
+3.3V
10kΩ
100nF
470Ω
100nF
4.7kΩ
MIXER ALERT LED2
A
K
λ
K
A
ZD1 1.8V
+1.8V
4.7kΩ
10µF
100nF
ZD1
K
A
100nF MKT
4.7kΩ
IC4: MCP6022
28 AVdd
13 Vdd
1 MCLR
2 RA0
18 RB9 SDA
17 RB8 SCL
RB12/DAC1RP 23
RB13/DAC1RN 24
RB14/DAC1LP 25
RB15/DAC1LN 26
4x 3.3kΩ
IC3 dsPIC33FJ128GP802
MISO1 15 RB6
MOSI1 16 RB7
SCK1 21 RB10
CS1 22 RB11
9 CLKI/OSC1/RA2
3 RA1
10 RA3/OSC2
12 RA4/SOSCO
RB0 4 MISO2
RB1 5 SCK2
AN4/RB2 6 MOSI2
PGED3/RB5 14
SOSCI/RB4 11
AN5/RB3 7
Vcap 20
Vss 8
Vss 19
AVss 27
10µF
IC4b
IC4a
1µF MKT
1µF MKT
4.7kΩ
100nF MKT
100nF MKT
4.7kΩ
4.7kΩ
100nF MKT
AUDIO OUT
CON2
CS2a
CS2b
CS2c
CS2d
CS2e
CS2f
MOSI2
SCK2
MISO2
MASTER CLK
+3.3V
CS2a
CS2b
CS2c
MOSI2
SCK2
MISO2
CS2d
CS2e
CS2f
TO IC8
TO IC9
100nF
10kΩ
28 AVdd
13 Vdd
1 MCLR
2 RA0
3 RA1
14 RB5
15 RB6
5 RB1
6 RB2/AN4
7 RB3/AN5
9 OSC1/CLKI/RA2
20 Vcap
SS1/RB0 4
SDI1/RB9 18
SDO1/RB8 17
SCK1/RB7 16
OSC2/RA3 10
RB4 11
RA4 12
RB10 21
RB11 22
DAC1RP/RB12 23
DAC1RN/RB13 24
DAC1LP/RB14 25
DAC1LN/RB15 26
IC10 dsPIC33EP512MC502
10µF
Vss 8
Vss 19
AVss 27
NOTE: IC7, IC8 AND IC9 NOT SHOWN, FOR CLARITY. BUT THEY ARE ALL CONNECTED IN EXACTLY THE SAME WAY AS IC5, IC6 AND IC10 EXCEPT FOR THE CS2d-f LINES SHOWN ABOVE.

Fig.8: the entire Synthesiser circuit, which is somewhat unusual in that it mainly consists of eight PIC microcontrollers (of three varieties), all communicating via two separate SPI serial buses. The remainder of the circuit comprises the EEPROM (IC11) used to store patch and performance data, the power supply, MIDI input and audio output.

mainly exists to limit the current into that pin and ensure that it's pulled to 0V when no USB connection is present.

IC2 and IC3 communicate via a second separate SPI bus, with a dedicated chip select line, from pin 7 of IC2 to pin 22 of IC3. IC2 also drives LED1, the MIDI Alert LED, from its RB6 digital output (pin 27).

External potentiometer VR1 (the volume control) connects to CON4, placing it across the 3.3V supply. Its wiper goes to analogue input AN11 of IC2 (pin 25). IC2 reads the voltage at the wiper using its analogue-to-digital converter (ADC) and passes the digital value along to IC3, which then scales its output to provide the desired volume level.



The Spectral Sound Synthesiser PCB is relatively easy to build. Although with about a dozen ICs, many of them having 28 pins, there are lots of solder joints to make. Be careful to make each joint properly or it won't function correctly.

The pot value or type isn't critical, but 100kΩ is reasonable. Scaling the audio sample values entering the DAC, rather than directly adjusting the op-amp gain, simplifies the PCB at the cost of reduced audio bit-resolution with the volume turned down. In practice, it's hard to hear this degradation.

That just leaves the MIDI input, clock signal distribution and the power supply to describe.

The MIDI signal is applied to CON6, and it powers the IR LED within FOD260L opto-isolator OPTO1. A 220Ω resistor provides current limiting, while diode D2 prevents the LED from being reverse-biased. It is essential to use the FOD260L opto-coupler as this is suitable for 3.3V operation – other varieties may well not work.

The output transistor in OPTO1 is operated in common-emitter mode with a 470Ω pull-up resistor. The resulting signal goes to the RX input (pin 18) of IC2.

A single external oscillator is used because we have eight microcontrollers that all need clock sources. This is built using crystal X1, its two 33pF load capacitors and unbuffered inverter IC1a. The resulting 16MHz signal is inverted by IC1b and buffered by IC1c and IC1d, then fed to all the microcontrollers' clock input pins.

We don't recommend using a buffered inverter in place of IC1, such as the more common 74HC04, as it might not oscillate correctly.

The power supply is simple; the unit is powered with 5V DC from barrel socket CON1, and this flows via a reverse-polarity protection diode to the inputs of linear regulators REG1 and REG2. REG1 powers all the digital circuitry while REG2 powers the analogue circuitry, which is basically just op amp IC4 and the bias for zener diode ZD1.

Increasing the signal-to-noise ratio (SNR)

A challenge with any system comprising mixed digital and analogue circuitry is to stop the digital noise bleeding through into the audio output. The module PCB takes the basic steps of separating audio and digital components as much as possible, with separate regulators and the use of a ground plane. However, additional measures have been taken to ensure a generally quiet and acceptable audio system.

One such measure is an audio limiter in the mixer audio code using advanced look-ahead AGC. A limiter squashes the dynamic range slightly by attenuating peaks, thereby effectively boosting the quieter sounds and lowering the noise floor.

This does not seem natural when trying to emulate polyphonic instruments; however, limiters and compressors are commonplace in audio reproduction, and it has a significant beneficial effect in this system.

We are also using a trick called pre-emphasis and de-emphasis. The digital audio generated has high-frequency boost applied, and the analogue signal processing circuitry has a matching high-frequency attenuation applied through a low-pass filter on the op amps. This way, the higher, more noticeable element of circuit noise is suppressed.

The module actually 'boosts' the higher harmonic levels by carefully attenuating lower harmonic levels. It is nice that complex digital filters are not needed to do this!

Finally, the Patch Editor application automatically boosts harmonic levels to the maximum, ensuring that the summed wavetable waveform is across the full signed 16-bit range, maximising the SNR.

Construction

The *Spectral Sound Synthesiser* is relatively straightforward to build, as the use of numerous microcontrollers minimises the number of separate components required. Most components mount on a double-sided PCB coded 01106221 that measures 145 x 94mm. The overlay diagram for this PCB is shown in Fig.9.

There is nothing remarkable about construction except that it requires good soldering skills to solder 200+

Programming the microcontrollers

This project uses eight microcontrollers of three different types. They are all Microchip products (one PIC and seven dsPICs), so they can be programmed with a PICkit 3, PICkit 4, Snap programmer or similar. Or you can build it from our kit, which will come with all the micros pre-programmed.

Each different type of micro has its own software. In other words, there are three sets of firmware. The codes are given in the parts list, and the download package on our website includes the source code for all three, plus the three HEX files you need to program them.

If you want to rebuild the source code to produce new HEX files (eg, you want to make changes to the way it functions), you'll need the Microchip XC16 Pro compiler (which can be ordered from the Microchip website; there are also free trial versions). Otherwise, optimisation level 3 will not be available, and the resulting firmware will not be fast enough to work correctly.

The PIC18 code is less critical, so you can probably get away with using the free XC8 compiler to build that HEX file.

pins accurately! We recommend using IC sockets throughout, including for the opto-coupler; while sockets can cause long-term reliability problems due to oxidation of the contact points, there is no real provision for in-circuit programming.

Still, since most constructors will be using pre-programmed micros (or programming them before assembly), you could consider soldering them directly to the board as long as you are confident they have been programmed correctly. Note that we haven't specified a socket for IC1 as there's little reason to use one there.

Start with the resistors, checking each lot of values with an ohmmeter before soldering them in place. Follow with the three diodes. Each is a different type, so don't get them mixed up, and ensure they are fitted with the cathode stripes facing as shown.

Next come the IC and opto sockets (or ICs and opto-coupler). Ensure they all have pin 1 facing towards the top of the board, and if soldering the ICs to the board, be very careful not to get the different 28-pin types mixed up.

After that, mount all the non-polarised ceramic and MKT capacitors; there are 100nF ceramic and MKT capacitors, so make sure the MKTs go in the positions shown in Fig.9.

Now install the electrolytic capacitors with the longer positive leads to the pads marked '+' in Fig.9, followed by the polarised pin headers and jack socket CON2.

Next, solder the LEDs in place with the longer leads to the side marked A. Fit these with sufficient lead length so that they will reach the top panel of the case once the PCB has been installed (see the section 'Wiring it up' below for a discussion on case selection). Follow with the two regulators, first attaching the heatsink to REG1 using the machine screw, nut and washer.

That just leaves crystal X1, DC socket CON1, USB socket CON5 and MIDI socket CON6 on the PCB. Mount those in order of increasing height. Finally, if you've soldered sockets to the board, plug in all the ICs and the opto-coupler now, paying careful attention to their pin 1 orientation and not getting the different 28-pin and 8-pin ICs mixed up.

Case selection

The PCB is designed to fit into the case specified in the parts list, and the front panel label (Fig.10), lid artwork (Fig.11) and drilling template (Fig.12) all fit that case. These can also be downloaded as PDFs and a PNG from the July 2023 page of the *PE* website at: **https://bit.ly/pe-downloads**

Parts List – Spectral Sound MIDI Synthesiser

1 double-sided PCB coded 01106221, 145 x 94mm from the *PE PCB Service*
1 instrument case **[Takachi YM-150; RS Cat 373-2255]**
4 stick-on rubber feet
1 front panel label, 145 x 37mm (see Fig.10)
1 lid label, 141 x 85mm (see Fig.11)
1 5-6V DC 1A regulated plugpack •
1 16MHz crystal, HC-49 (regular or low-profile) (X1)
1 PCB-mount DC barrel socket, 2.1mm or 2.5mm ID to suit plugpack (CON1)
1 3.5mm stereo DPST switched jack socket (CON2) **[Altronics P0094, RS Cat 913-1021 or CUI SJ1-3555NG]**
1 2-pin polarised header and matching plug (CON3, for power switch)
1 3-pin polarised header and matching plug (CON4, for volume control)
1 through-hole full-size type-B USB socket (CON5) **[Jaycar PS0920, Altronics P1304A/P1304B]**
1 5-pin 180° DIN socket, right-angle PCB mount (CON6) **[Jaycar PS0350, Altronics P1188B or RS Cat 491-087]**
1 SPST or SPDT panel-mount slide switch (S1, power)
1 100kΩ panel-mount linear potentiometer and knob (VR1, volume control)
8 28-pin narrow DIL IC sockets (**optional**; for IC2, IC3 and IC5-IC10)
3 8-pin DIL IC socket (**optional**; for IC4, IC11 and OPTO1)
1 TO-220 heatsink (REG1) **[maximum 40mm wide, 13mm deep from tab, <18°C/W; RS Cat 263-251 used for prototype]**
4 M3-tapped 6.3mm spacers
1 M3 x 10mm panhead machine screw, shakeproof washer and hex nut
8 M3 x 5mm panhead machine screws
2 M2 x 10mm countersunk screws and nuts (for slide switch mounting)
1 100mm length of rainbow cable (for wiring to S1 and VR1)
1 small tube of thermal paste
• *up to 9V can be used, but 5-6V results in more reasonable dissipation*

Semiconductors
1 74HCU04 unbuffered hex inverter, DIP-14 (IC1)
1 PIC18LF25K50-I/SP 8-bit micro programmed with 0110622A.HEX (IC2)
1 dsPIC33FJ128GP802-I/SP 16-bit microcontroller programmed with 0110622B.HEX (IC3)
1 MCP6022-I/P rail-to-rail op amp, DIP-8 (IC4)
6 dsPIC33EP512MC502-I/SP 16-bit microcontrollers programmed with 0110622C.HEX (IC5-IC10)
1 24LC512-I/P 64Kbyte I^2C EEPROM, DIP-8 (IC11)
1 FOD260L opto-coupler, DIP-8 (OPTO1)
2 LF33CV 3.3V low-dropout linear regulators (REG1, REG2)
2 3mm high-brightness green LEDs (LED1, LED2)
1 1.8V 250mW zener diode (ZD1) **[eg, 1N4614]**
1 1N4004 400V 1A diode (D1)
1 1N4148 75V 150mA signal diode (D2)

Capacitors
1 100µF 6.3V electrolytic
1 10µF 16V electrolytic
8 10µF 16V X7R ceramic
2 1µF 63V MKT
4 100nF 63V MKT
13 100nF 50V X7R ceramic
2 33pF 50V ceramic

Resistors *(all ¼W 1% metal film axial)*

1 1MΩ	6 4.7kΩ	1 1kΩ	1 100kΩ	4 3.3kΩ
2 470Ω	10 10kΩ	4 2.2kΩ	1 220Ω	

UK builders – Kit SC6261 – We recommend UK builders purchase an almost complete kit from Silicon Chip. It includes programmed dsPICs and everything else except the case, feet, labels and plugpack. AU$260 *including* delivery to the UK. See: https://www.siliconchip.com.au/Shop/20/6261

You could use a different case provided it's large enough to house the PCB, since all the connectors/controls (apart from the two which are panel-mounted) are along one edge of the PCB.

With a 145 x 94mm PCB, most cases measuring at least 165 x 100mm should be suitable. The height required depends on the heatsink you are using for REG1. The specified

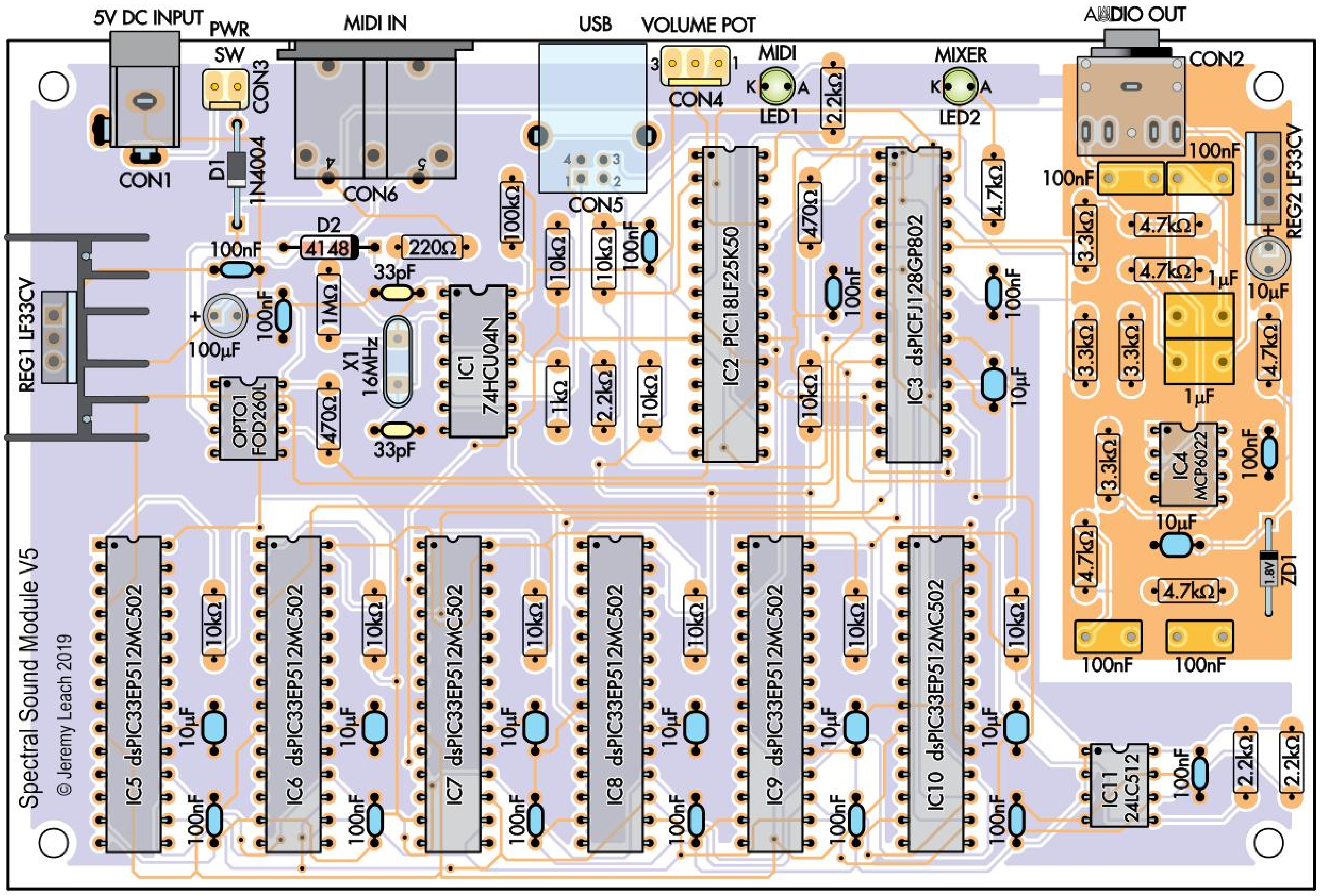

5V DC INPUT
PWR SW
CON3
MIDI IN
USB
VOLUME POT
AUDIO OUT
CON2
CON1
D1 1N4004
CON6
CON5
CON4
MIDI
LED1
MIXER
LED2
D2 4148
100nF
220Ω
100kΩ
10kΩ
33pF
1MΩ
X1 16MHz
IC1 74HCU04N
IC2 PIC18LF25K50
IC3 dsPICFJ128GP802
REG1 LF33CV
REG2 LF33CV
100μF
OPTO1 FOD260L
470Ω
1kΩ
2.2kΩ
4.7kΩ
3.3kΩ
1μF
10μF
IC4 MCP6022
ZD1
IC5 dsPIC33EP512MC502
IC6 dsPIC33EP512MC502
IC7 dsPIC33EP512MC502
IC8 dsPIC33EP512MC502
IC9 dsPIC33EP512MC502
IC10 dsPIC33EP512MC502
IC11 24LC512
Spectral Sound Module V5
© Jeremy Leach 2019

Fig.9: like the circuit diagram, the eight PICs dominate the overlay, all in 28-pin DIL packages. Make sure to orient those correctly and don't get them mixed up.

Fig.11: the lid panel artwork (shown at approximately 85% actual size) is a nice finishing touch to the project. It's designed to be printed onto a transparent medium. Note that the two LED positions could vary somewhat, especially if you're using a different case; you could simply cut that part of the decal off and position it separately. ►

heatsink is only 20mm tall, so cases at least 35mm tall should be fine. If you're using a taller heatsink, add 10-15mm to its height to figure out what cases will be suitable.

Possible alternative instrument cases include Altronics Cat H0374 or Cat H0378 (with a short heatsink), Jaycar Cat HB5912 or the Hammond RM2055M, which is available from Digi-Key and Mouser. It should also fit into a UB2 Jiffy box like Jaycar Cat HB6012, Altronics Cat H0152 or H0202, but they don't look as good as instrument cases, and it will be a bit harder to fit the board.

Mount the board in the case using machine screws and tapped spacers so that its top edge is against one side of the case (for an instrument case, it should be the front panel). Next, mark and drill/cut holes in the adjacent panel for the DC power barrel plug, MIDI input socket, USB socket and audio output jack.

If you're using the specified case, you can use the drilling diagram (Fig.12) to assist you. It could also be used on other cases, but you will need to adjust the placement on the panel to match your PCB mounting location.

Wiring it up

Once you've confirmed these are all accessible through the panel, if you haven't already, drill holes for the volume pot and power switch in convenient locations. Then solder appropriate lengths of ribbon cable strips to those parts and crimp/solder pins to the other ends that you then push into the plastic polarised header blocks (or solder direct to the PCB).

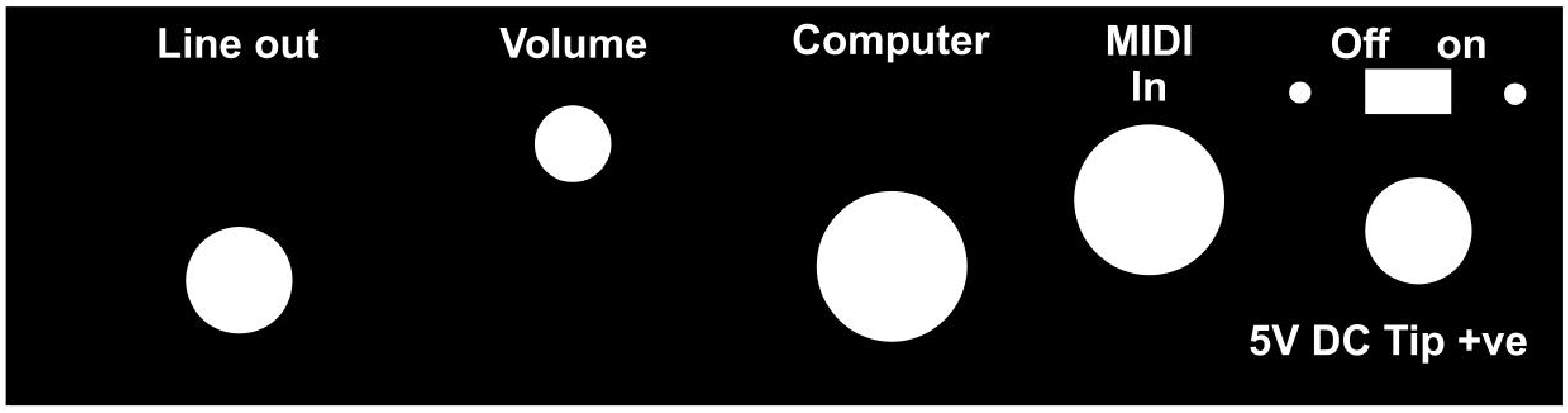

Line out
Volume
Computer
MIDI In
Off on
5V DC Tip +ve

Fig.10: the front panel artwork can be downloaded, printed, laminated (or protected in another manner) and then attached to the drilled panel.

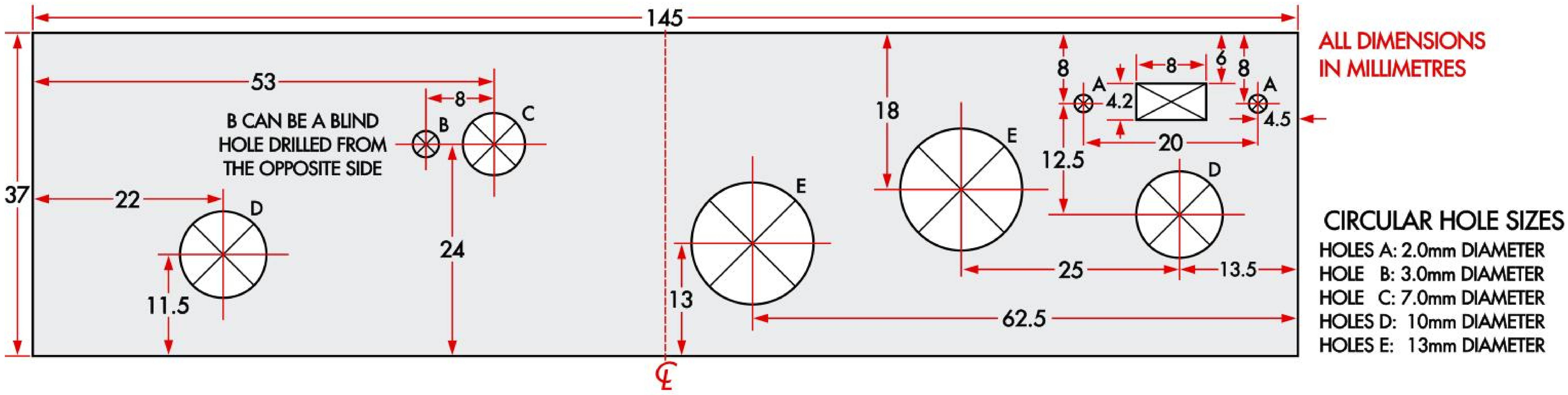
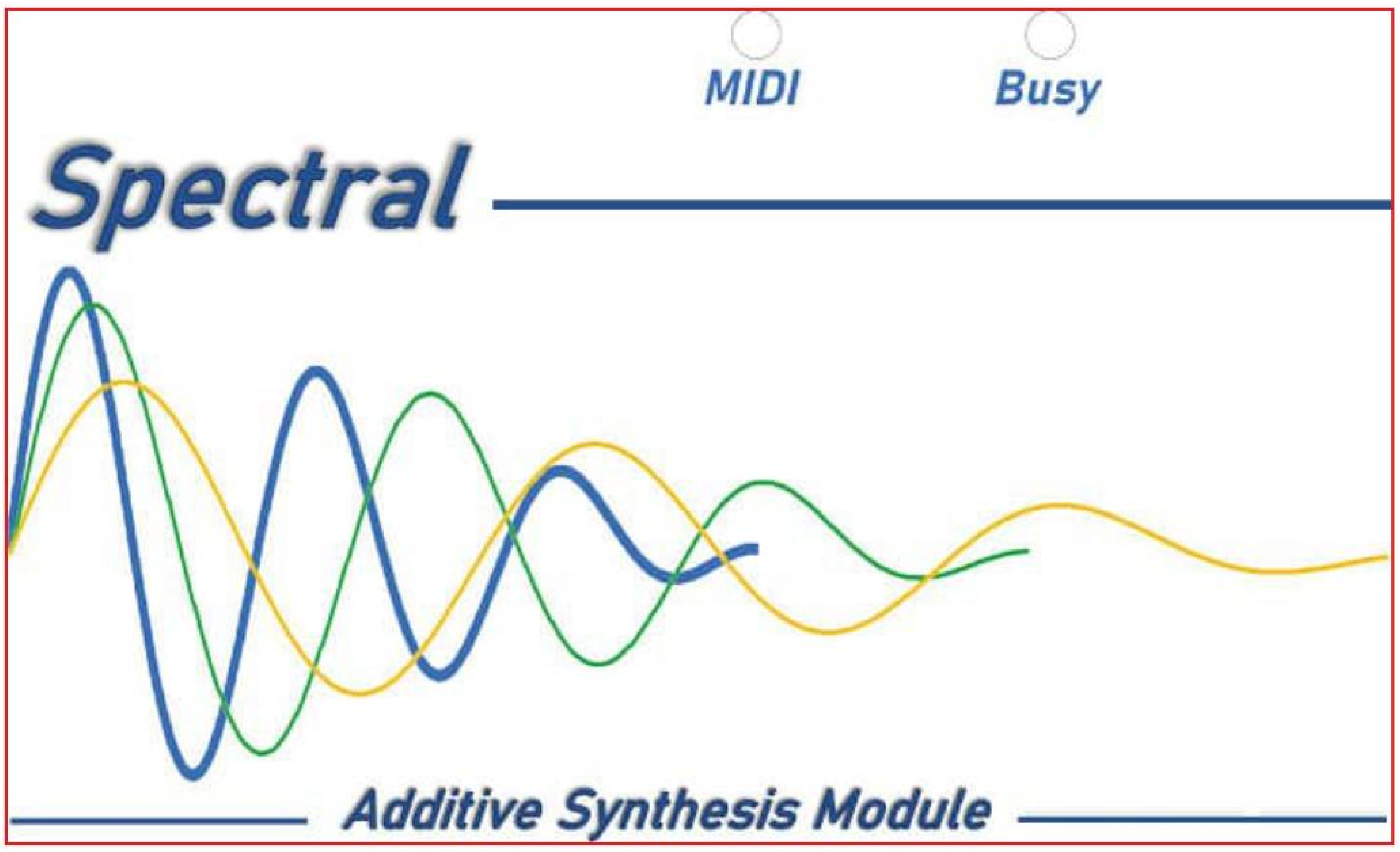

145
53
8
B CAN BE A BLIND HOLE DRILLED FROM THE OPPOSITE SIDE
B
C
37
22
D
24
11.5
E
13
℄
18
E
8
A
4.2
8
6
8
A
4.5
20
12.5
D
25
13.5
62.5
ALL DIMENSIONS IN MILLIMETRES
CIRCULAR HOLE SIZES
HOLES A: 2.0mm DIAMETER
HOLE B: 3.0mm DIAMETER
HOLE C: 7.0mm DIAMETER
HOLES D: 10mm DIAMETER
HOLES E: 13mm DIAMETER

Fig.12: the positions of the holes to drill/cut in the front panel. The volume control and on/off switch are panel-mounted, so they could be moved, but these positions are designed to clear other nearby parts. You can use this for cases other than the recommended one, but you'll need at least one reference point to position it correctly.



You will also need to mark and drill two 5mm holes in the lid for the LEDs to protrude through. Depending on their lead lengths, you might have a little bit of flexibility in where those LEDs are placed as you can bend the leads slightly. Keep in mind that if you are applying the lid panel label, it will have to line up with those holes.

Now is a good time to adhere the front and/or lid panel labels (see below for hints on making them) and cut out the holes using a sharp hobby knife.

Verify that S1 and VR1 are wired up correctly, mount them on the front panel and then plug them into headers CON3 and CON4. You can then 'button up' the board inside the case, power it up and check that it's operational.

To do that, you will need to plug it into a computer running Windows, download and install the software described in the following section, and verify that it can connect to the *Spectral Sound Synthesiser*.

Making the labels

I've found that just printing with an inkjet printer and then spraying with art 'fixing' spray works well. For the lid label, I use decal sheets for my inkjet printer. They have a paper backing. The process is:

1. Print just like you would to any sheet of paper (but with the shiny side up).
2. Spray with varnish/lacquer/fixative.
3. Submerse in water for 30 seconds to a minute.
4. Gently slide the decal off (the very thin decal detaches from the paper backing when wet) and onto the case.
5. Consider varnishing the dried decal for added protection.

The 'Patch Editor' software

The module has an associated, powerful Windows program called the 'Patch Editor', written in C# Winforms. A screengrab of this software is shown in Screens 1 and 2.

This is a 'Click-Once' .NET application that I am hosting online at: **https://bit.ly/pe-jul23-patch**

This is in Microsoft Azure 'blob' storage, which means that the user is notified of version enhancements if installed from this online location. A comprehensive user guide for this software is available.

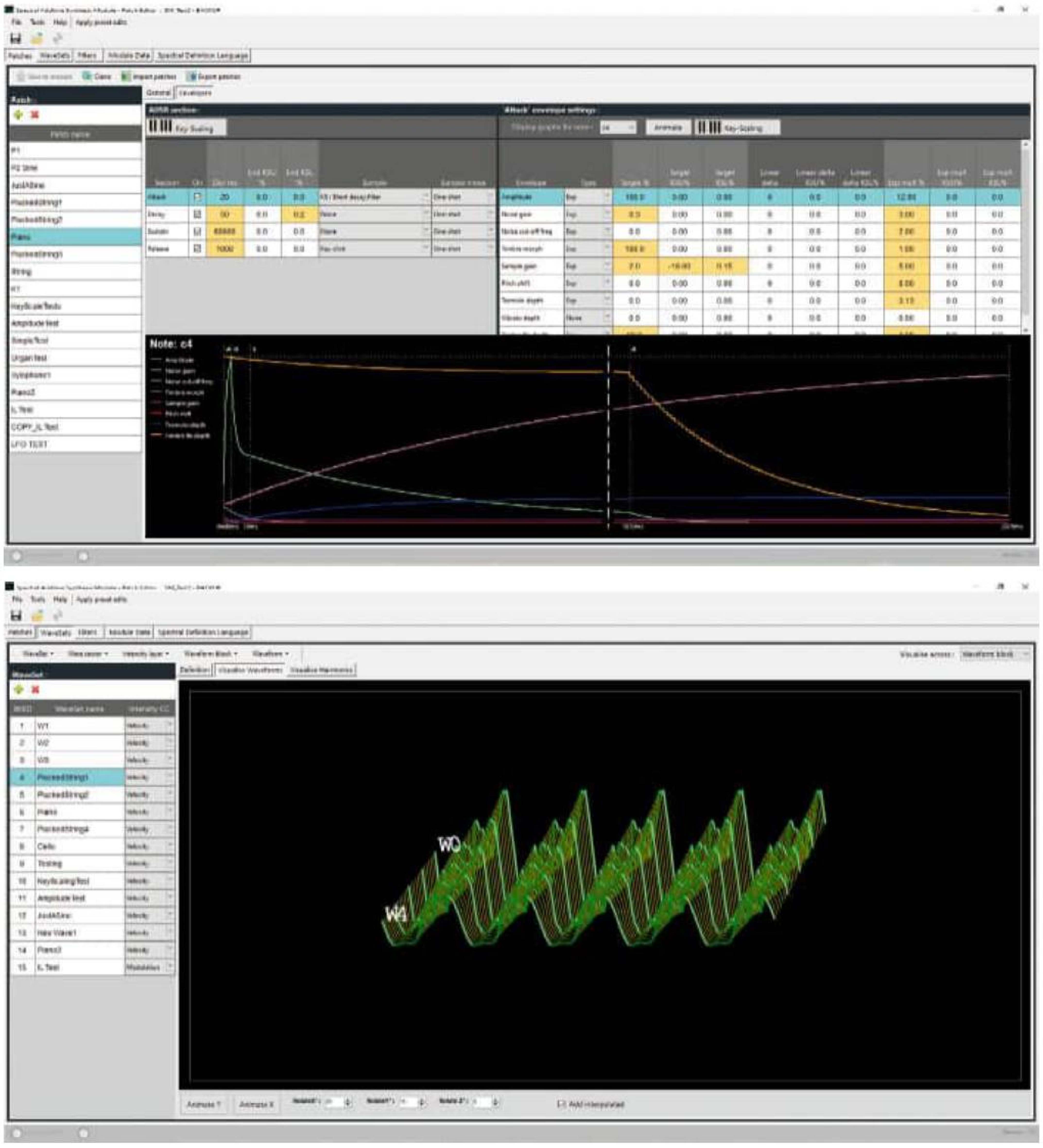
The app includes tools to help shape the timbre 'landscape', the envelopes, the filters and more. It includes 'visualisers' to view the timbres both in the time or frequency domain, and even includes a harmonic analyser, where you can grab the harmonic content from audio!

The app also has its own unique programming language called 'Spectral Definition Language' (SDL) *[not to be confused with Simple DirectMedia Layer – Editor]*. You can write SDL code to finely tune the patch definitions and easily reuse chunks of code. The idea is to 'abstract' sound design to a higher level, simplifying all the complexities of detailed configuration.

To this end, you can store your own code snippets and execute them as necessary via the app menu – a powerful concept, with 'out of the box' default examples for setting things like a 'Hammered String' envelope!

Final thoughts

This project has been a very intense but rewarding journey, often feeling like it is 'shooting for the moon'. It shows that sound synthesis is still fertile soil for experimentation and invention. Fig.13



Screens 1 and 2: sample screenshots from the powerful Windows-based Patch Editor software designed to interface with the Spectral Sound Synthesiser. Its source code is included in the download package.

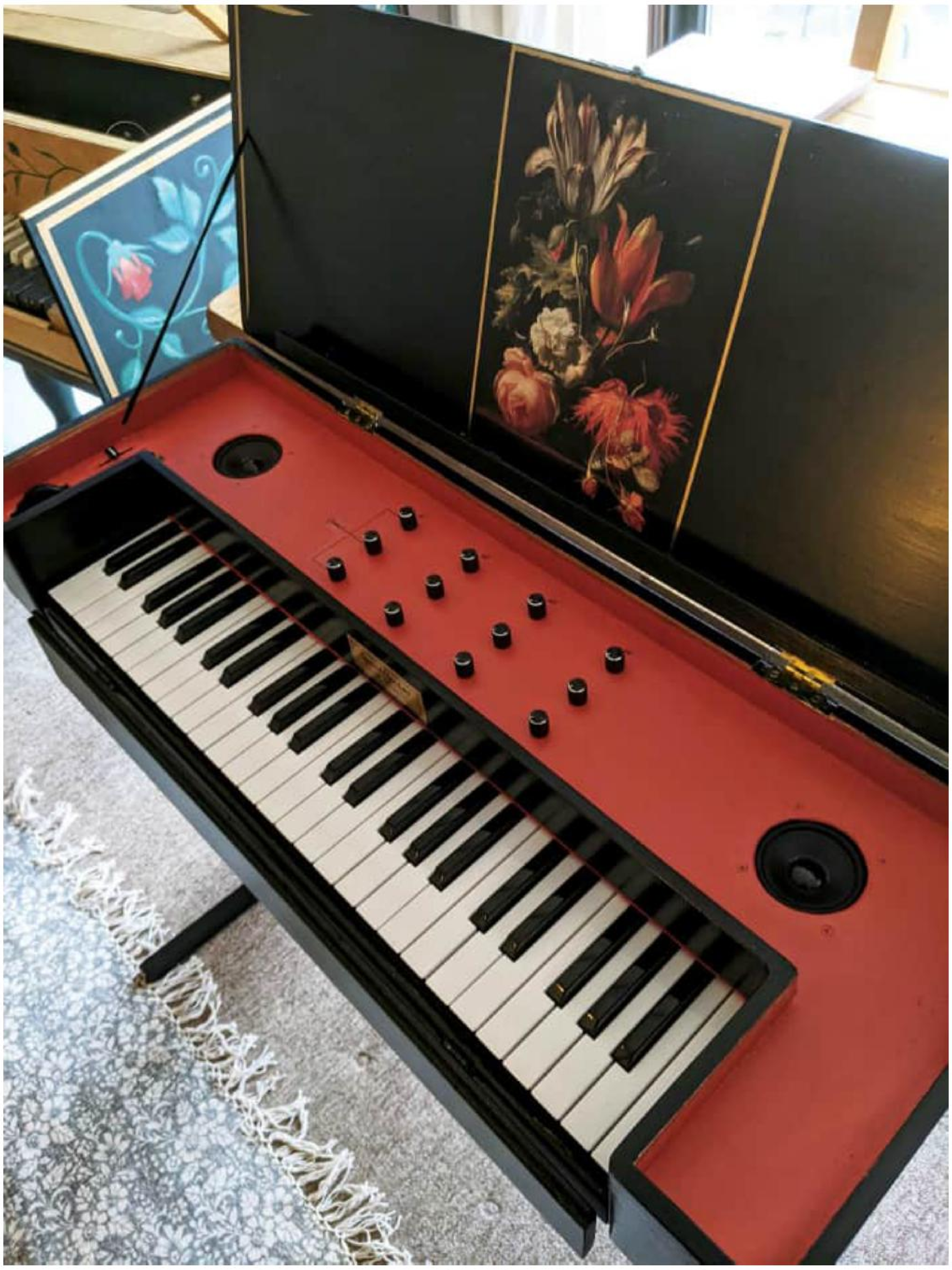


Fig.13: the MIDI Synthesiser was combined with a standard MIDI controller keyboard, amplifier and speaker to form this electronic clavichord.

shows my DIY 'Electronic Clavichord' containing a standard MIDI controller keyboard coupled with this module and a tiny amplifier and speaker.

One tantalising idea for the future could be to approach the problem of a timbral-based system from a more holistic angle. Rather than each played note having its own wavetable, think of the required harmonics from all played notes as one giant pool of oscillators. We could then use the phenomenon of 'psychoacoustic masking' to significantly 'prune' the actual harmonics that require calculation.

This would require the ability to prioritise the harmonic importance and ignore the ones of least significance. An interesting aspect of this approach would be that the threshold could be based on system performance, always processing the maximum number of harmonics possible but degrading the sound quality in a controlled way if needed.

This approach might also be able to deviate from the integer-based harmonic requirement, thus offering more realism.

Another idea is to return to a more sample-based approach, but instead of storing samples in the conventional PCM way, keep them as timbres or even as timbre changes. This might provide significant savings.

Other ideas start questioning fundamentals about the precision needed for harmonic levels. Since humans perceive sound logarithmically, adequate level scaling might result from simple bit shifting. Can we really tell the difference in harmonic levels to such a degree that it justifies anything better?

Moreover, we need to think more about how our brains perceive sound, and less about the purity of mathematical calculation. Our brains work on impression and recognising overall characteristics, so maybe there's potential in focusing on techniques that make huge computational savings by disregarding things that just do not matter to perception. It seems there is still much to think about regarding sound synthesis!

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Useful Links

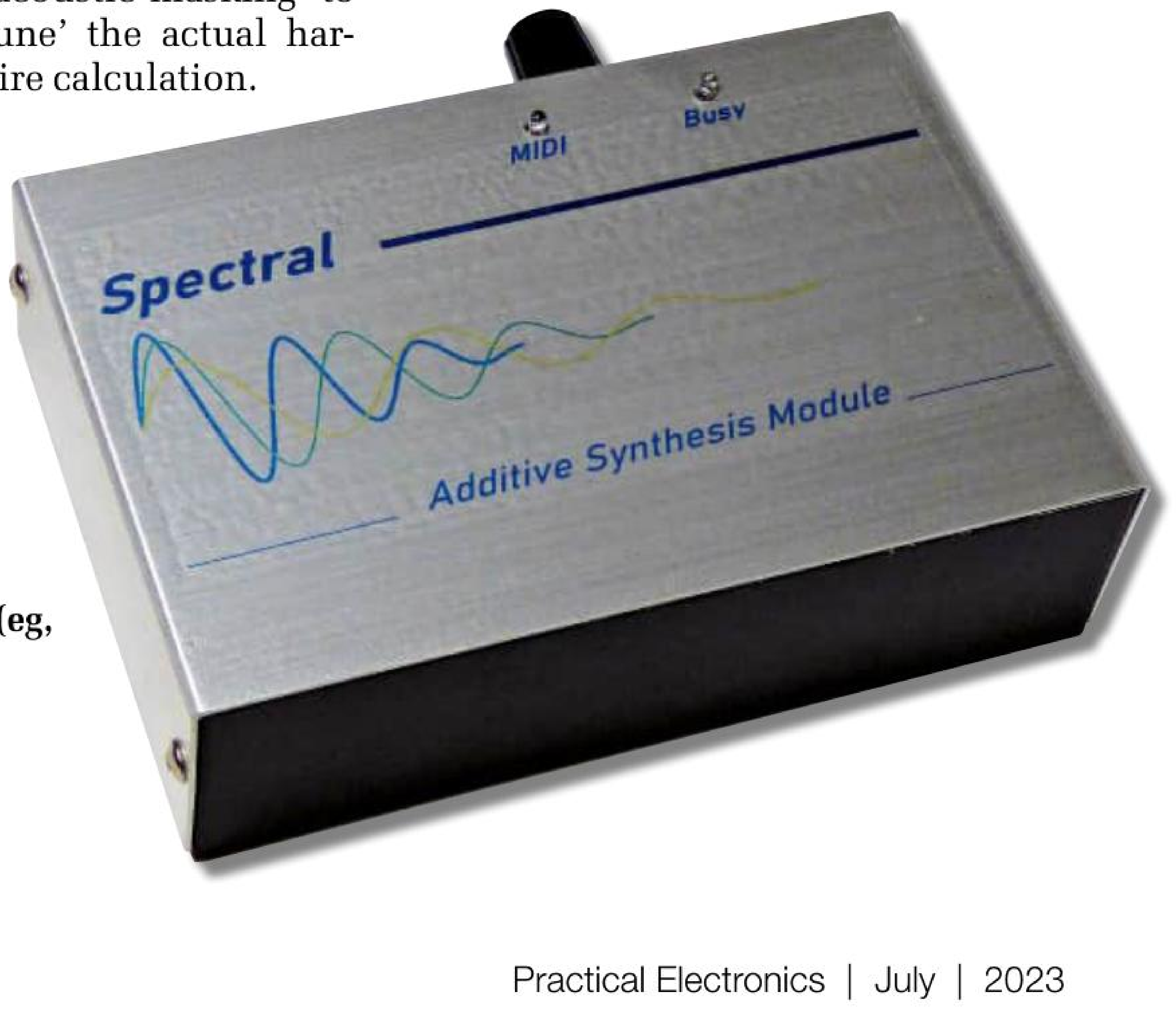
The biological bases of musical timbre perception:
https://bit.ly/pe-jul23-bio

Synthesising plucked strings:
https://bit.ly/pe-jul23-pluck

Synthesising wind instruments:
https://bit.ly/pe-jul23-wind

Sound quality or timbre:
https://bit.ly/pe-jul23-tim1

Details on timbre:
https://bit.ly/pe-jul23-tim2



The finished Synthesiser has two LEDs on the top of the panel to indicate when it is receiving MIDI messages and when it is communicating with a computer (eg, loading patches).

MULTIMETER -CHECKER -CALIBRATOR

215
200
2k
20k
200M
2M
20M
200M
200m
V
2
20
200
1000
750
200
20
2
200m
A
2m
20m
200m
20
20
200m
20m
2m
2000
200
2
200
20
hFE
COM
V

It’s amazing how handy even the cheapest multimeters can be – but, are they accurate? This Multimeter Checker can verify their accuracy, and for meters that aren’t so cheap, it will also allow you to calibrate them and adjust for drift.

Project by Tim Blythman

Multimeters are indispensable tools; perhaps so necessary that we tend to take them, and their accuracy, for granted. Sometimes accuracy is not that important, but there are times when it is.

Back in August 2016, we presented the *Low-cost Accurate Voltage Current Resistance Reference* for checking and calibrating multimeters.

It provides a DC reference voltage of 2.5V ±1mV (±0.04%), a resistance of 1kΩ ±1Ω (±0.1%) and a current of 2.5mA ±3.5μA (±0.14%). The DC voltage reference comes from a precision voltage reference IC, and that plus a precision resistor provides the current reference. That precision resistor can also be used on its own as the resistance reference.

The whole thing is compact and ran from a coin cell, perfect for keeping in the toolbox to be used whenever needed. It covers the most common measurements done with a multimeter.

While that was great, it didn’t provide an AC voltage source, so not all of the typical multimeter ranges could be checked/calibrated. So, we decided to develop a new design to add that feature.

For the new *Multimeter Checker*, we have a dedicated voltage reference IC providing 3.3V for DC calibration. This is also used with a precision resistor to provide an accurate 100mA current source. It has another precision resistor to act as a resistance reference.

Importantly, for calibrating AC voltage ranges, it provides a precise 1V RMS AC sinewave at one of three frequencies: 50Hz or 60Hz (to match typical mains frequencies) or 100Hz.

Different multimeters use different methods to measure AC voltage (and alternating current). That is why some multimeters are labelled as ‘True RMS’ while others are not.

True RMS multimeters give accurate AC voltage measurements, whatever the shape of the waveform. In contrast, some cheaper multimeters measure the peak voltage and multiply the reading by a factor of 0.71, on the assumption that the waveform is sinusoidal. Of course, this will not be accurate unless the waveform is close to being a sinewave.

A square wave, for example, will give an artificially low reading as its peak is the same as its RMS value. Similarly, triangle and sawtooth waves will tend to give readings that are too high.

Some other meters measure the average of the rectified AC voltage and assume a sinewave, which will have different error magnitudes for other waveforms.

In our circuit, the AC voltage is generated by an analogue circuit, so it does not have the digital artefacts that would be produced by a digital synthesis method. Its amplitude and frequency are checked and adjusted by a microcontroller, which compares these to the DC voltage reference and the frequency of a crystal oscillator.

As well as enjoying the benefits of both analogue and digital circuitry, this allows the AC voltage reference to be set to 50Hz, 60Hz or 100Hz.

Circuit details

The entire *Multimeter Checker/Calibrator* circuit is shown in Fig.1. The DC references (voltage, current and resistance) on the *Checker* work much the same as in the earlier *Low-cost Accurate Voltage Current Resistance Reference*. Still, we’ll explain how they work together, because they are also an intrinsic part of the AC voltage reference.

5V USB power is applied to socket CON1 and powers, among other things, 3.3V precision voltage reference VREF1. This MCP1501 low-cost 3.3V precision reference is critical to the correct operation of all the other parts.

It’s capable of supplying up to 20mA, which is vital to ensure that the accuracy

Features and specifications

- **DC voltage reference........................3.3V ±0.1%**
- **AC voltage reference........................1V ±0.5% RMS**
- **Direct current reference.................100mA ±0.2%**
- **Resistance reference ....33Ω ±0.1%**
- **AC reference voltage frequency.... 50Hz, 60Hz or 100Hz (±0.3%)**
- **AC reference frequency source..... crystal oscillator**
- **AC reference harmonics .... ≤40dBV**
- **Control via pushbuttons with LEDs, and over USB virtual serial port**

of the reference is not affected by the connected loads, especially as the analogue generation circuitry is powered from this 3.3V reference.

VREF1 has a 100nF bypass capacitor at its positive supply, pin 1. The 3.3V output from pin 7 is connected to TP5 and can be compared with circuit ground at TP6; these two points are marked DCV on the PCB. The PCB has separate circuit traces from TP5 to REF1's feedback (FB) pin 8. This ensures that the 3.3V is accurate at the test point, in spite of any loads.

Precision reference current

Dual low-voltage rail-to-rail op amp IC3 (MCP6272) is powered from the 3.3V output of VREF1 and has a 100nF supply bypass capacitor. One half of IC3 (IC3b) is used to drive the current reference.

The 3.3V from VREF1 feeds into the non-inverting input of IC3b (pin 5) via a 1kΩ resistor. Its corresponding inverting input (pin 6) is fed (via another 1kΩ resistor) from the high side of a 33Ω precision resistor used to measure the reference current. Any current through this resistor causes a voltage to develop between TP4 and ground.

The output of this op amp (pin 7) drives the base of NPN transistor Q1, acting as an emitter-follower, via a 100Ω resistor. Q1's collector is connected to the 5V rail, and its emitter goes to TP3.

TP3 and TP4 are thus the current reference terminals. When TP4 is below 3.3V, Q1 is fed current by the op amp. If TP4 starts to rise above 3.3V, the current drive to Q1 starts to get cut off. When TP4 is at 3.3V, 100mA must be flowing through the 33Ω resistor to ground.

There will be a minuscule current flowing from TP4 into the op amp's pin 6, but it is of the order nanoamps, so it is much less significant than the 0.1% precision component tolerances.

Thus, the op amp's feedback loop maintains 100mA between TP3 and TP4 when the two are connected by a multimeter measuring current. TP3 and TP4 are labelled on the PCB as the DCA reference points.

The 1nF capacitor between pins 6 and 7 helps suppress any high-frequency oscillation that might occur due to the high gain of the op amp.

With 3.3V across the 33Ω resistor plus the base-emitter drop of Q1 and perhaps 0.1V across the 100Ω base resistor, the op amp output is typically at 4V, giving about 1V of headroom below the 5V supply. So anything connected to the current reference must drop less than 1V or have less than 10Ω resistance for the current reference to work correctly.

A second, identical 33Ω precision resistor is provided as the resistance reference, allowing the circuit to provide an independent set of test pads, TP7 and TP8, for the resistance feature.

AC voltage reference

Practically all of the remaining circuitry is used to provide the AC reference.

Since this circuit operates from a single-ended 5V DC supply, we first need a nominal level around which the AC signal can swing. For this, we have chosen half of the 3.3V supply, which is derived by using a pair of 10kΩ resistors to divide the output from VREF1 to produce 1.65V.

The resulting voltage is low-pass filtered by a 1μF capacitor and buffered by IC3a, with another 10kΩ resistor providing the unity-gain feedback. The output of this op amp (pin 1) sits at 1.65V, and this is our AREF rail.

The AC signal is generated by a phase-shift oscillator based around another op amp, IC1 (another MCP6272), and IC2, an AD8403ARZ10 quad 10kΩ digital potentiometer.

IC2 is powered by the 5V rail with a 100nF bypass capacitor. The analogue ground pins 1, 5, 17 and 21 connect to circuit ground, along with its digital ground at pin 9, while the $\overline{\text{SHDN}}$ (shutdown) and $\overline{\text{RS}}$ (reset) pins are pulled up to 5V by 10kΩ resistors to allow normal operation of the digital potentiometer at all times.

Op amp IC1 is powered from the 3.3V rail, with a 100nF bypass capacitor, to provide signal symmetry around the 1.65V AREF reference. This is one reason why we have chosen the MCP1501 reference, as it has a sufficient output current and suitable voltage to power these components.

This is critical because one of IC1's outputs saturates briefly on every cycle, so if it were powered from 5V, the saturation would occur differently on positive and negative swings, leading to harmonics (ie, frequencies above the selected 50/60/100Hz option) creeping into the output.

A phase-shift oscillator works by reinforcing a signal that is delayed by 360°. The delay is formed by several RC filter networks, which add up to 180° of phase shift, followed by inversion, equivalent to a further 180° phase shift.

As the RC filter phase shift depends on frequency, it will only have a delay of precisely 360° at one specific frequency. Signal components at other frequencies are attenuated as they are delayed by a different amount and interfere destructively as they make their way around the circuit.

The circuit elements also attenuate all frequencies to some extent, so one half of op amp IC1 provides the gain needed to overcome this, while the other half provides the phase inversion.

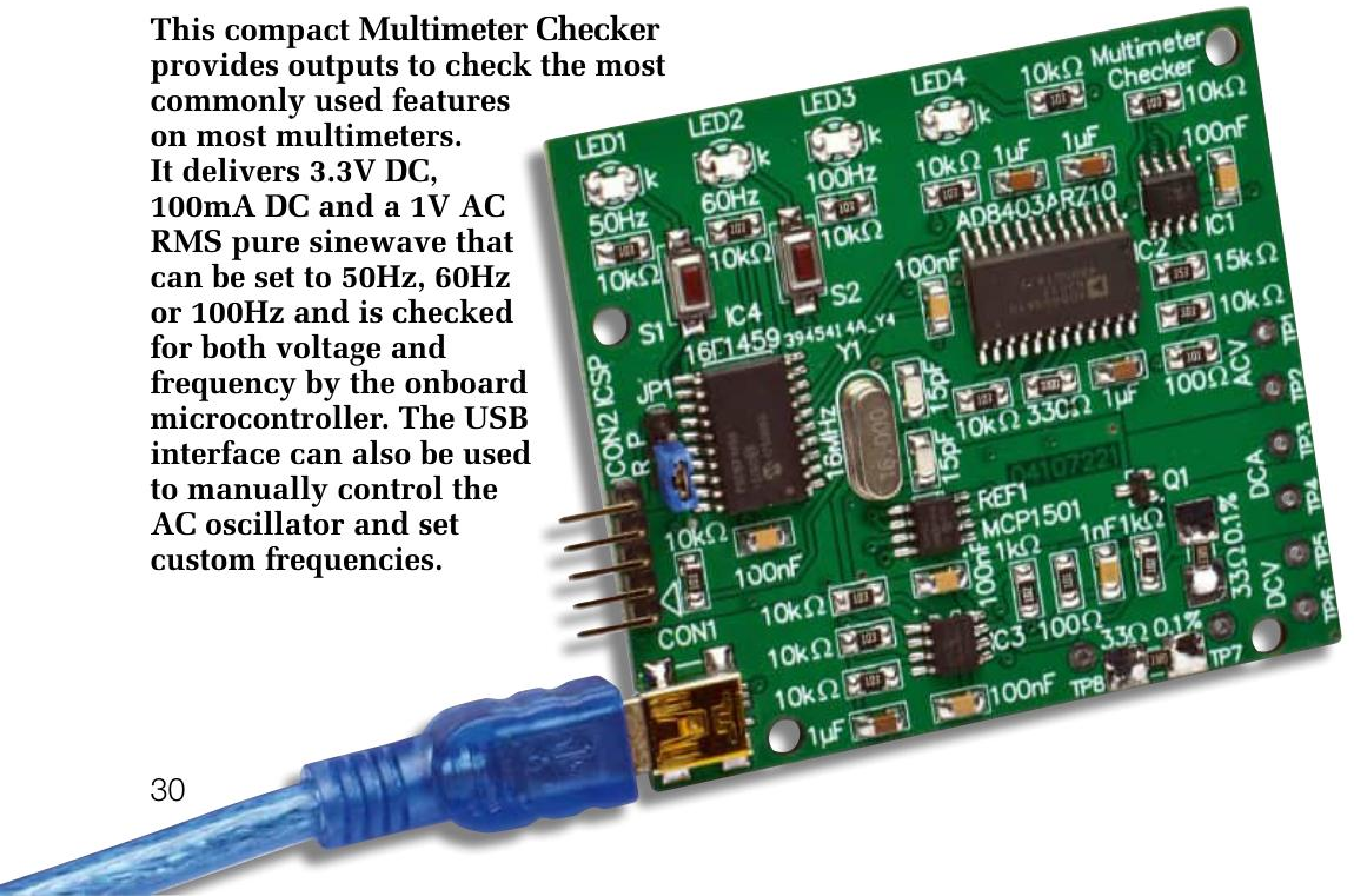
Phase shift oscillator

There are three phase-shift elements composed of three 1μF capacitors connected to IC2 and three of the digital potentiometer elements inside IC2 (numbered 1-3). These are all wired as variable resistors (rheostats) and can vary independently from near to 0Ω up to around 10kΩ.

Imagine a fairly pure 50Hz 1V AC RMS signal at pin 1 of IC1; this is what is expected when the oscillator is working as designed and set to the 50Hz output. 1V RMS is around 2.8V peak-to-peak.

Op amp IC1b acts as an inverting amplifier with a gain of 1.5 times. So the output at pin 7 is expected to be an inverted version of IC1a's pin 1 signal, but with a 4.2V peak-to-peak value. Since IC1 is fed from a 3.3V supply, the output saturates at 3.3V peak-to-peak.

The resulting waveform is between a sinewave and a square wave, so it will also have some odd harmonics of 50Hz present, the first of which is at 150Hz. Fig.2 shows the spectrum of the oscillator's output at 50Hz. Note that the only significant harmonic is the third at 150Hz, although its level is down by over 40dB compared to the fundamental.



This compact Multimeter Checker provides outputs to check the most commonly used features on most multimeters. It delivers 3.3V DC, 100mA DC and a 1V AC RMS pure sinewave that can be set to 50Hz, 60Hz or 100Hz and is checked for both voltage and frequency by the onboard microcontroller. The USB interface can also be used to manually control the AC oscillator and set custom frequencies.

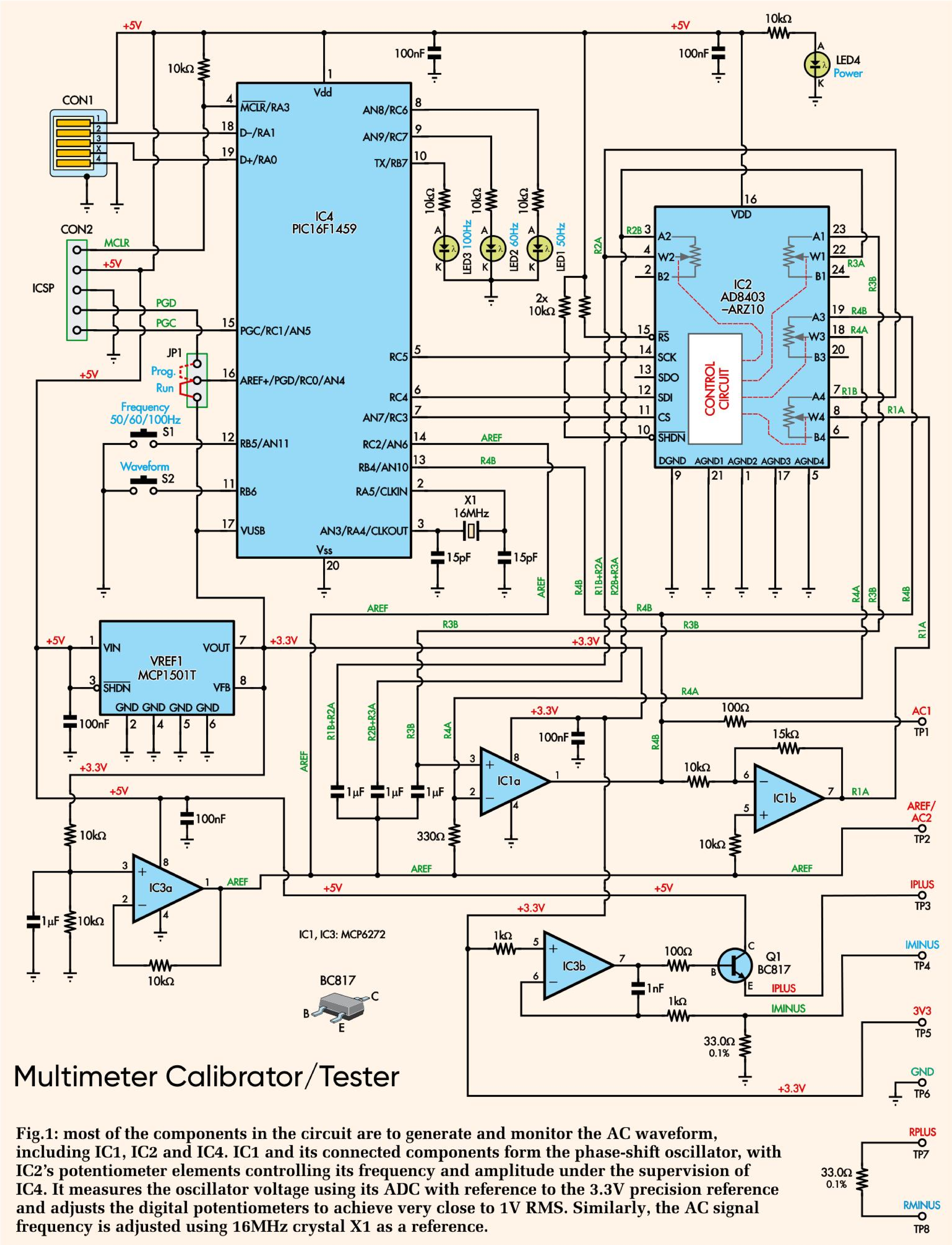

Multimeter Calibrator/Tester
CON1
CON2
ICSP
MCLR
+5V
PGD
PGC
JP1
Prog.
Run
Frequency 50/60/100Hz
S1
Waveform
S2
IC4
PIC16F1459
Vdd
MCLR/RA3
D–/RA1
D+/RA0
PGC/RC1/AN5
AREF+/PGD/RC0/AN4
RB5/AN11
RB6
VUSB
Vss
AN8/RC6
AN9/RC7
TX/RB7
RC5
RC4
AN7/RC3
RC2/AN6
RB4/AN10
RA5/CLKIN
AN3/RA4/CLKOUT
X1
16MHz
15pF
100nF
10kΩ
LED3 100Hz
LED2 60Hz
LED1 50Hz
LED4
Power
2x 10kΩ
IC2
AD8403
–ARZ10
CONTROL CIRCUIT
VDD
A2
W2
B2
A1
W1
B1
A3
W3
B3
A4
W4
B4
RS
SCK
SDO
SDI
CS
SHDN
DGND
AGND1
AGND2
AGND3
AGND4
R2A
R2B
R3A
R3B
R4A
R4B
R1A
R1B
R1B+R2A
R2B+R3A
AREF
+3.3V
VREF1
MCP1501T
VIN
VOUT
VFB
GND GND GND GND
1μF
330Ω
IC1a
IC1b
100Ω
15kΩ
AC1
TP1
AREF/ AC2
TP2
IPLUS
TP3
IMINUS
TP4
3V3
TP5
GND
TP6
RPLUS
TP7
RMINUS
TP8
IC3a
IC3b
IC1, IC3: MCP6272
BC817
B
C
E
1kΩ
1nF
Q1
BC817
33.0Ω
0.1%

Fig.1: most of the components in the circuit are to generate and monitor the AC waveform, including IC1, IC2 and IC4. IC1 and its connected components form the phase-shift oscillator, with IC2's potentiometer elements controlling its frequency and amplitude under the supervision of IC4. It measures the oscillator voltage using its ADC with reference to the 3.3V precision reference and adjusts the digital potentiometers to achieve very close to 1V RMS. Similarly, the AC signal frequency is adjusted using 16MHz crystal X1 as a reference.

Note also that we will still obtain a 3.3V peak-to-peak output from IC1b even if the signal that comes from IC1a's pin 1 output drops as low as around 0.8V AC RMS, or if it was higher than 1V AC RMS due to the previously described saturation effect.

This amplified signal from IC1b (pin 7) passes through the three RC low-pass filters. If the digital pots are set to around 5.5kΩ, each stage will cause a 60° delay to the 50Hz component and approximately halve its AC amplitude (as measured at each successive capacitor).

Other, higher-frequency components will be delayed more and attenuated even more. For example, the third harmonic of 50Hz at 150Hz will be phase-shifted by around 80° and be reduced to about a fifth of its original amplitude by each stage.

The three stages interact to a degree, so a simple mathematical analysis of each stage separately does not quite match what happens when they are combined. Before building the prototype, we had to simulate the entire circuit to determine the required component values.

The result is a relatively pure 50Hz signal, but with quite a low amplitude coming into pin 3 of IC1a. But as long as the pin 7 output of IC1b is saturated on each cycle, the level is steady.

IC1a acts as a non-inverting amplifier with a gain set by the ratio of the 330Ω fixed resistor and the fourth variable resistor in IC4. This gain is selected to bring the attenuated signal from the RC filter stages back up to 1V RMS and is fed to TP1 via a 100Ω resistor to protect IC1 from external short circuits.

TP2 is connected to the 1.65V reference so that the sinewave between TP1 and TP2 can be measured without a DC offset.

So, the AC signal frequency can be changed by adjusting the three variable resistor elements in the three RC networks. Similarly, the amplitude can be varied by adjusting the fourth variable resistor value.

The resulting waveforms are shown in Scope 1. The primary output signal is the blue trace, while the red trace is the saturated output at IC1b's pin 7. Note that it is inverted compared to the blue trace.

You can see that the orange, yellow and green traces are phase-shifted and attenuated by each successive RC stage. The green trace is amplified to become the blue trace, thus completing the feedback loop.

Control circuitry

IC4 is a PIC16F1459 microcontroller that adjusts and monitors the AC reference for accuracy, among other tasks. It is powered from the 5V USB supply with a 100nF bypass capacitor between pin 1 (5V) and pin 20 (ground). A 10kΩ resistor between pins 1 and 4 pulls up the $\overline{\text{MCLR}}$ pin to allow normal operation when the circuit is powered.

IC4 needs both an accurate voltage and frequency reference to do its job. The 3.3V output of VREF1 goes to JP1, and with the appropriate jumper fitted (in the 'Run' position), it feeds through to pin 16 (AREF+) of IC4. Since pin 16 also provides the PGD programming function, JP1's other jumper position (marked 'Prog.') connects to programming header CON2.

The other programming signals from IC4 are also connected to CON2. This includes $\overline{\text{MCLR}}$, 5V, ground and PGC at IC4's pin 15.

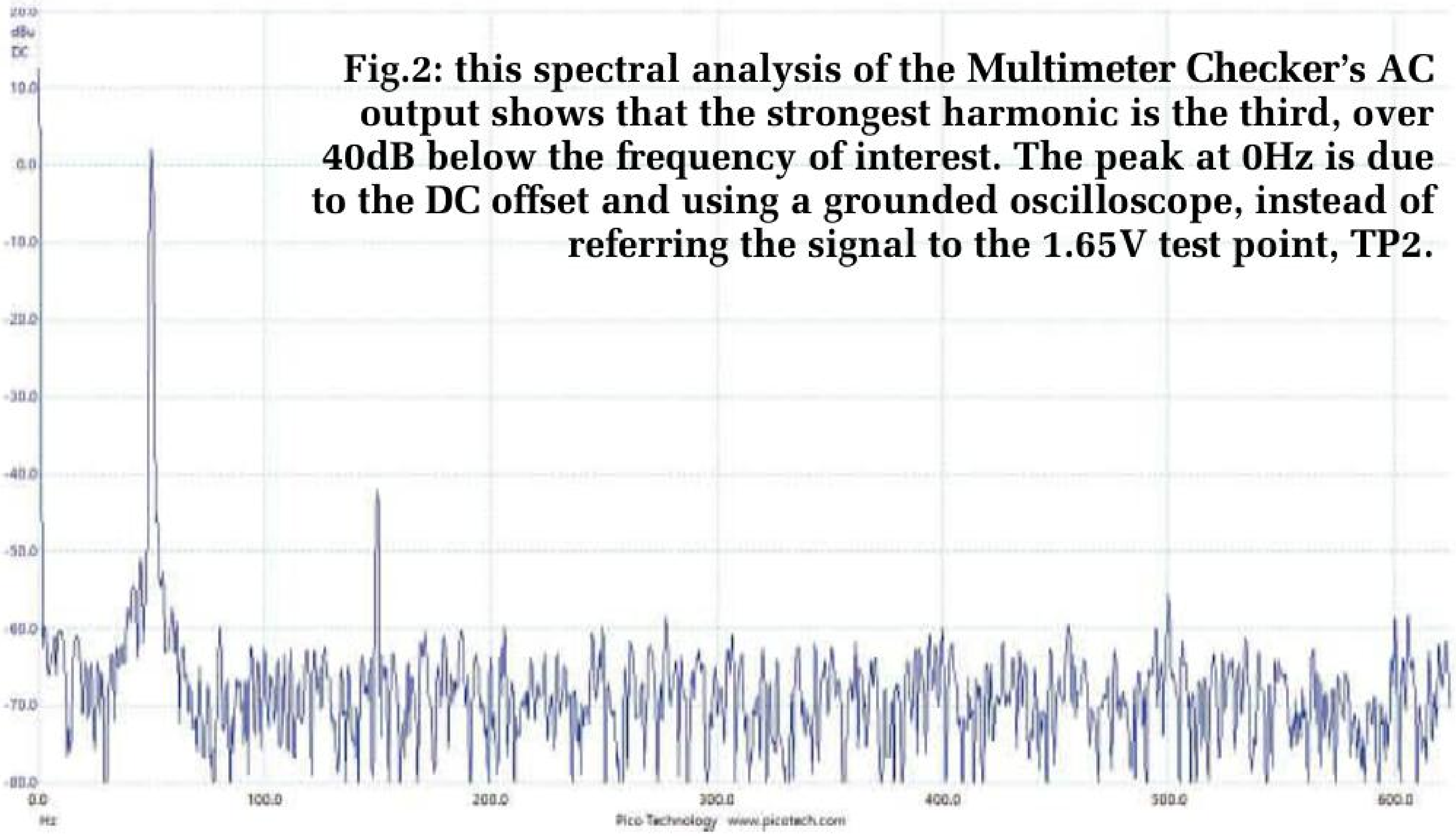


Fig.2: this spectral analysis of the Multimeter Checker's AC output shows that the strongest harmonic is the third, over 40dB below the frequency of interest. The peak at 0Hz is due to the DC offset and using a grounded oscilloscope, instead of referring the signal to the 1.65V test point, TP2.

Pins 13 and 14 connect to the AC reference output at TP1 and the 1.65V AREF signal, respectively. These are monitored by the ADC (analogue-to-digital converter) peripheral in IC4 to check the frequency and amplitude of the output signal.

The frequency reference comes from 16MHz crystal X1, connected to IC4's pins 2 and 3 (CLKIN and CLKOUT). A 15pF load capacitor connects from each side of the crystal to circuit ground so it will oscillate correctly.

Three LEDs (LED1-LED3) connect to IC4 via 10kΩ series resistors. The LED cathodes are grounded, so the LEDs illuminate when pins 8-10 are pulled high.

Two tactile pushbuttons, S1 and S2, connect to pins 11 and 12. The other side of each switch is grounded while the pins are internally pulled up, allowing the micro to detect when the button is pressed. These LEDs and buttons provide a basic control interface for operating the *Multimeter Checker*.

Control of digital potentiometer IC2 is over an SPI serial interface, with pins 5, 6 and 7 of IC4 being connected to pins 14, 12 and 11 of IC2. These lines have the roles of SCK (clock), SDI (data) and $\overline{\text{CS}}$ (chip select), respectively.

Since IC2 uses an unusual 10-bit interface and a high data rate is not needed, the SPI commands are sent via bit-banged GPI/O operations. This also allowed us to simplify the PCB layout as we did not need to use the dedicated SPI (MSSP peripheral) pins, but could use any digital I/O pins.

Pins 17, 18 and 19 are associated with IC4's USB peripheral, so pins 18 and 19 are taken to the CON1 USB socket, and pin 17 is fed 3.3V from REF1. This means that the *Multimeter Checker* can be controlled and monitored by being connected to a computer's USB port too.

The PIC16F1459 was chosen as a suitable part because we could not quite fit the necessary features onto a 14-pin microcontroller. But the presence of the USB interface means that we can add some other interesting and valuable features too.

Finally, we get to the power supply. We've chosen a USB supply for its ubiquity. The 5V supply also gives more headroom than the 3V coin cell from the earlier design. After all, the 3.3V voltage reference would not function from a 3V cell. It also allows us to produce a higher test current than a coin cell could supply.

LED4 and a 10kΩ series resistor are connected across the incoming 5V supply to show that power is present. There is no onboard 5V regulator; we rely on the USB source to be within the normal 4.5-5.5V range. All of the onboard components running from the 5V rail can handle that.

Firmware

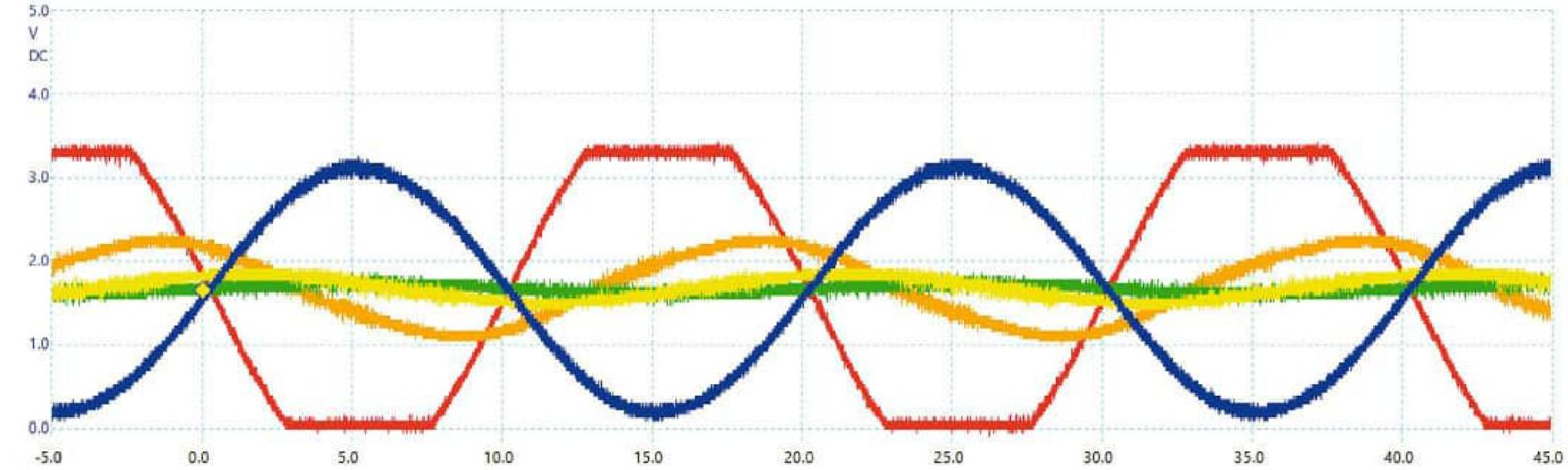
The firmware program that runs on IC4 has three main aspects. The first is the fairly straightforward task of monitoring the buttons S1 and S2 and controlling LEDs LED1-LED3, providing a basic user interface.

The second is the USB interface. This appears as a virtual serial port when connected to a computer. Keystrokes from the computer are stored in a buffer and handled much like button presses, but with extra functions.

There is also the option of 'printing' status updates to the serial port, so that the *Multimeter Checker* can provide more detailed information via the virtual serial port than can be displayed with the LEDs.

Finally, IC4 is responsible for setting and monitoring the AC reference voltage output. It has no control over the DC voltage or current references, although it uses the DC voltage reference to check the AC voltage. The crystal oscillator used for IC4's timebase ensures that all timing is accurate, particularly in measuring the frequency.

The microcontroller samples the AC voltage waveform and checks its period (and thus its frequency),



Scope 1: the blue trace is the AC output signal at TP1, while the red trace is measured at output pin 7 of IC1b. The orange, yellow and green traces are measured at the top of each 1µF capacitor to the left of IC1a in Fig.1, from left to right.

peak-to-peak amplitude and average absolute amplitude (with reference to the 1.65V midpoint).

Since the 3.3V reference is used as the scale for the ADC peripheral, the absolute digital value of the peak-to-peak and average amplitude values are known and fixed in the program.

The sampling works as follows. A timer interrupt fires 6000 times every second and takes a sample of the AC waveform. We chose this rate to allow integer divisions of 50Hz, 60Hz and 100Hz into that timer. Although that is not critical, it makes the calculations simpler.

Just over 240 samples are taken, corresponding to two complete cycles at 50Hz. This is so that we can ensure that at least two positive-going zero crossings occur within each sample set; these are the points between which the period is measured.

While 120 samples for a cycle at 50Hz does not seem like much precision, the firmware interpolates where the zero crossings occur to within 1/16th of a sample. It does this by calculating how much the samples before and after the zero-crossing are above or below the zero point. This way, the period can be measured with a resolution of around one part in 960 for a 100Hz signal, or better for lower frequencies.

Sampling must occur without interruption, so a set of samples is taken and then processed. Adjustments are made if necessary; then it goes back to sampling. By taking both the peak-to-peak and average amplitude, the Checker can also confirm that the waveform is sinusoidal, as a waveform with a different shape will not be able to match both.

Oscillator adjustments

The four digital potentiometers each have 256 steps. This is what limits the amplitude accuracy to 0.5% (about 1 part in 200), as the steps are about that far apart.

In practice, a small amount of dithering occurs, so the average over several cycles will be closer to the target, close to the accuracy of the 3.3V reference.

The frequency can be controlled more closely than the amplitude, as three potentiometers are involved. Rather than stepping all three together, each is incremented in turn, giving almost three times as many steps.

This resolution results in steps of around 0.1Hz at 50Hz up to 0.3Hz at 100Hz, around 0.3% in the operating range. Like the amplitude, dither over several cycles improves the longer-term average accuracy of the frequency.

We'll mention the full details of the USB interface a bit later. It provides a manual mode that allows direct control of the digital potentiometers.

Construction

The *Multimeter Checker* is built on a small PCB, 65 × 58.5mm, coded 04107221 and available from the *PE PCB Service* – see Fig.3. It is mainly populated with surface mounting parts, although they are all pretty large and easy to work with. The only part with a smaller pin pitch than 1.27mm is the USB socket, and all passives are M3216/1206 parts at around 3.2 × 1.6mm.

We'll assume you have flux, solder wick, tweezers and all the other gear for working with these sorts of parts. Fume extraction is a good idea when working with flux.

Start by fitting the USB socket, CON1. Apply flux to the pads on the PCB and insert the socket's locating posts into their holes on the PCB. Clean the iron's tip and add fresh solder. Carefully apply the tip to each lead in turn without touching the metal shell.

After soldering each pin, use a magnifier to check that there are no solder bridges, and if there are, use the wick to remove them. If you can't see, clean off the flux residue with alcohol or a flux cleaner.

If you find a solder bridge, apply fresh flux to the leads and press the wick against the bridge using the iron, then carefully pull both away. When the smaller leads look tidy, solder the larger pads for the shell, turning up the heat if necessary.

Fit the four ICs and REF1 next. These are all SOIC (small outline IC) parts of various sizes, but don't mix up REF1, IC1 and IC3 as they all have eight pins. Note too that IC3 and REF1 face in opposite directions. Check the part markings against the parts list and PCB silkscreen as you go, making double sure that pin 1 is correctly oriented in each case before soldering any pins.

For each part, apply flux, then tack one lead in place, ensuring the correct orientation by checking the silkscreen dot and IC markings. If the pads are all well aligned, solder the remaining pins; otherwise, adjust as needed by reapplying heat from the iron.

Like with CON1, check for solder bridges and remove them as needed. It's usually easier to solder all the pins before removing any bridges.

Q1 is the only transistor on the board, and it should be fitted as shown in the photos and overlay. It's the smallest part overall, so be careful not to lose it. But as the leads are widely spaced, it should not be difficult to solder.

Install the capacitors next. The values will not be marked on the parts themselves, so work with one value at a time. The values required for each location are shown in Fig.3.

Solder one lead, check that the part is square, flat and even within its pads and then solder the remaining lead. Refresh the first lead if necessary. Remember to add flux to the PCB pads as you go, regularly cleaning the iron tip and then adding fresh solder.

The resistors should be marked with codes representing their values. They are all the same size; check Fig.3 or the PCB silkscreen to see which values go where.

We used larger pads for the 33Ω precision resistor in case part shortages meant that we couldn't get the high-accuracy parts in an M3216/1206 size, so don't be concerned that the part is much smaller than the pads.

Now fit the four LEDs. They are all in one corner of the PCB and have their cathodes to the right, as indicated by the cathode symbol on the silkscreen. You can use either M3216/1206 surface-mounting types or 3mm through-hole LEDs.

For through-hole LEDs, the anode lead is usually longer. If using SMD LEDs, they should have green cathode markings, but it's pretty easy to check them with a DMM set on diode test mode.

Hold the probes on either side of the LED (making sure it doesn't fly away!). If the LED lights up, the red probe is on the anode and the black probe on the cathode.

The two tactile switches mount near the LEDs. Install these in the same fashion as the other two-lead parts. That completes the surface-mounted parts, so you can now clean off the flux residue. The remaining components are all through-hole types, and some are optional.

Fit crystal X1 next. You should not need an insulating pad under the metal case as the two mounting pads are covered with solder mask on the top of the PCB. However, if the solder mask in that area is damaged, add an insulator or mount it off the PCB surface. Regardless, verify after soldering it that its case is not shorted to either pad underneath.

If you have already programmed the microcontroller (IC4), you don't need to fit CON2, the in-circuit programming header. In this case, you could also replace JP1 with a short wire link across the pair of pads on the 'R' side of the jumper.

Otherwise, fit both headers and install the jumper shunt initially in the 'P' position for programming.

Although we have not used them on our prototype, we've scattered a few 3mm holes around the PCB to fit standoffs if you want to mount the *Multimeter Checker* to something.

Programming

If you haven't pre-programmed the microcontroller, you will have to program it now.

Using a PICkit 3, PICkit 4 or Snap connected to CON2, load the **0410722A.HEX** file onto IC4 using the Microchip IPE (integrated programming environment). If you are using a Snap, you likely will need to supply power to the board; this can be done using a USB lead connected to CON1.

When power is applied, LED4 will light up. So if you don't see LED4 illuminated, check for power and that the circuit has been built correctly before proceeding.

After programming, disconnect the programmer and move JP1 from the 'P' (program) position to the 'R' (run) position.

Testing

When the unit is powered up, it will start in 50Hz mode, and LED1 should be solidly lit to indicate this. Pressing S1 will cycle through the 50Hz, 60Hz and 100Hz modes. LED1-LED3 light up in turn to show the current mode.

Pressing S2 switches between the default pure sinewave to a more saturated waveform. You can use this facility to check how the multimeter responds to AC waveforms that are not pure sinewaves.

Parts List – Multimeter Checker and Calibrator

1 double-sided PCB coded 04107221, 65 × 58.5mm available from the *PE PCB Service*
1 mini USB Type B socket (CON1)
1 5-pin right-angle header (CON2; **optional**; only used for in-circuit programming)
1 3-pin header and jumper shunt (JP1)
2 small SMD two-pin tactile switches (S1, S2)
1 16MHz low-profile HC-49 crystal (X1)

Semiconductors
2 MCP6272 or MCP6L2 dual low-power rail-to-rail op amps, SOIC-8 (IC1, IC3)
1 AD8403ARZ10 4-channel 10kΩ digital potentiometer, wide SOIC-24 (IC2)
1 PIC16F1459-I/SO microcontroller programmed with 0410722A.HEX, wide SOIC-20 (IC4)
1 MCP1501T-33E/SN 3.3V voltage reference, SOIC-8 (REF1)
4 green LEDs, 3mm through-hole or M3216/1206 SMD (LED1-LED4)
1 BC817 50V 800mA NPN transistor, SOT-23 (Q1)

Capacitors *(all 10V+, X7R or C0G ceramic, SMD M3216/1206 or M2012/0805)*
4 1µF
5 100nF
1 1nF
2 15pF

Resistors *(all M3216/1206 1% ⅛W except as noted)*
1 15kΩ
12 10kΩ
2 1kΩ
1 330Ω
2 100Ω
2 33Ω 0.1%

In this mode, the amplitude is set to a high level (causing saturation of the op amp output and clipping). The LEDs indicate this mode by flickering rapidly. This waveform may be easier to verify during initial testing, as it does not depend on the microcontroller correctly detecting the amplitude.

If the LEDs are flashing slowly (around 1Hz), the *Multimeter Checker* has not been able to verify that the output frequency and amplitude are correct. They might flash briefly on a mode change, but there is a problem if they continue flashing for more than a few seconds.

In this case, first double-check that JP1 is in the run position. This connects the 3.3V reference to the microcontroller, so if it is still in the programming position or not fitted, the micro cannot confirm the AC output level.

One bad solder joint, especially around IC1 and IC2, will be enough to corrupt the waveform, so check those areas too.

If you have an oscilloscope, you can verify that the waveform at TP1 is a 1V RMS sinewave offset by 1.65V DC. The DC level can be eliminated by using AC coupling on the 'scope. Be careful not to ground TP2 unless the supply to the *Multimeter Checker* is floating (for example, it is powered by a USB battery pack).

USB control

Connecting the USB interface to a computer will provide a lot more information, so do this if possible, especially if you are troubleshooting. The *Multimeter Checker* should not need USB drivers on recent operating systems, and you can simply use a serial terminal program to communicate.

We usually use TeraTerm on Windows, but programs like Putty, the Arduino Serial Monitor or MMEdit can also be used. On Linux, minicom is one option.

Find out what serial port has been allocated and open this with your terminal program. You will not need to set a baud rate as it is a virtual serial port.

Typing '1', '2' or '3' will change the mode to 50Hz, 60Hz or 100Hz. You will see the LEDs change as the mode changes. Pressing 'S' selects the sinewave mode, while the 'R' key sets the saturated output (think 'rectangular wave').

Pressing the space bar will produce a status report over two lines; this can be seen at the top of Screen 1. The first line shows the current control variables; 'A' controls the amplitude and 'F' controls the frequency. The second line shows the reported amplitude (V) and frequency (F).

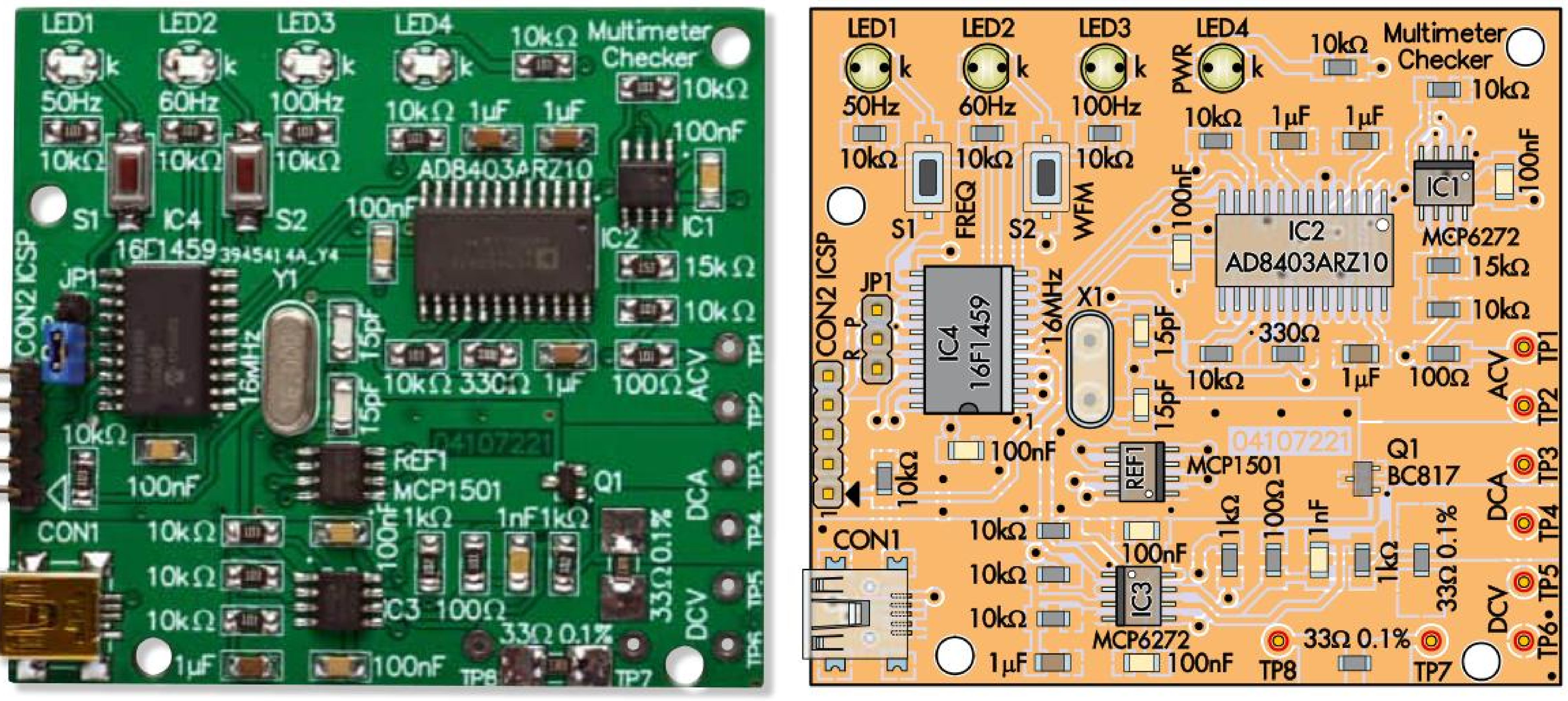


Fig.3: most components are relatively easy to solder; the USB socket is a bit tricky because its pins are pretty close together. During assembly, the most critical thing to check is that all ICs are oriented correctly, with their pin 1s in the positions shown. Also ensure that the solder makes contact with the pad and pin of each device and check carefully for solder bridges between pins when you've finished.

Pressing 'M' sets manual control mode. All three LEDs will light together in this case, and you can set the A and F parameters manually. The A parameter is changed with the full stop and comma keys (think of the <> above them on the keyboard). Increasing the A parameter will decrease the output amplitude.

Once the output voltage drops below 0.8V AC RMS, it may drop off altogether as there is insufficient gain around the feedback loop to maintain oscillation. Still, it will recover once a valid setting is selected.

You can change the frequency with the '−' and '+' (or '=') keys. The F parameter can span between 1 and 750, corresponding to approximately 45Hz to over 1kHz. The *Multimeter Checker* cannot accurately display frequencies over about 600Hz, so the use of this end of the range is not recommended.

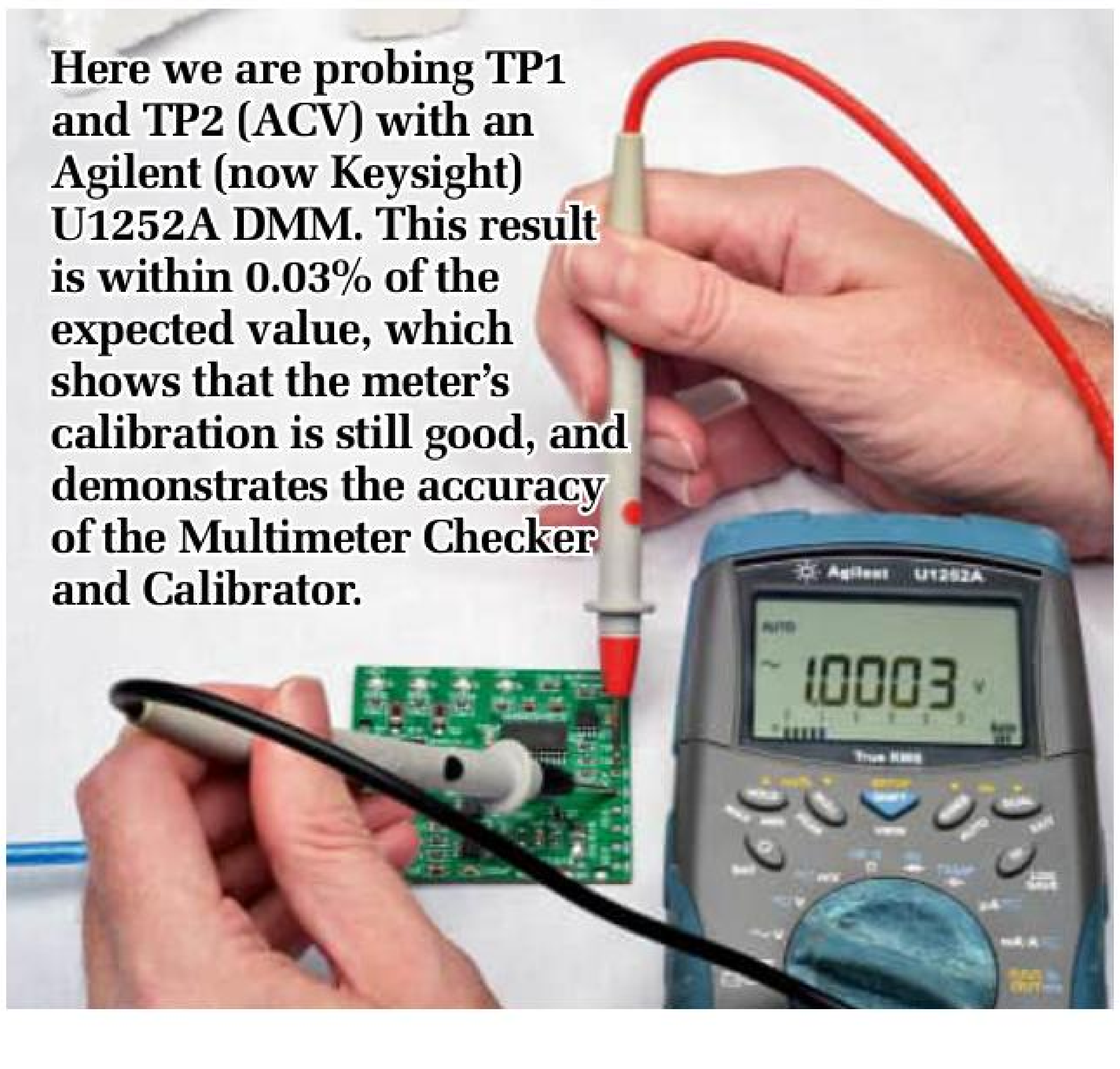
Manual mode is terminated by pressing S1 on the board, or selecting the 50Hz, 60Hz or 100Hz modes from the USB interface using the 1-3 keys.

Using it

Before you start using our *Multimeter Checker*, you should refer to the calibration section in its manual (if present).

When using our *Multimeter Checker*, you can check or calibrate a multimeter in the following modes:

- **DC voltage** – connect the probes between TP5 and TP6 on a range like 20V DC and check/adjust for 3.300V.
- **AC voltage** – connect the probes between TP1 and TP2 on a range like 2V AC and check/adjust for a reading of 1.00V. This should be correct regardless of whether the meter is a True RMS type or not, as it is a pure sinewave.
- **Direct current** – connect the probes between TP3 and TP4 on a range like 200mA and check/adjust for 100mA output. TP3 is the current source and TP4 is the sink, so you might get a negative reading unless the red probe goes to TP3.
- **Alternating current** – connect the probes between TP1 or TP2 with a 100Ω 1% or 0.1% resistor in series. Set it for a low range and check for a reading of 10mA.
- **Resistance** – connect the probes between TP7 and TP8 on a range like 200Ω and check/adjust for a reading of 33.00Ω.
- **Frequency** – connect the probes between TP1 and TP2 on a range like 200Hz and check for a reading of 50Hz, 60Hz or 100Hz (set using pushbutton S1 and LEDs1-3). For best results, press S1 until LED3 lights and check/adjust for 100.0Hz.
- **Duty cycle** – connect the probes between TP1 and TP2 and check for a reading of 50%. For best results, press S1 until LED3 lights.
- **True RMS readings** – press S2 to activate the modified wave mode and check the AC voltage reading between TP1 and TP2. The displayed voltage should be above 1V RMS; our prototype produces 1.27V RMS in this mode. A higher reading suggests your meter uses the average method. In comparison, a lower reading suggests the peak method (as the peak-to-peak voltage in this mode is 3.3V, a peak-reading multimeter will generally show around 1.17V).



Here we are probing TP1 and TP2 (ACV) with an Agilent (now Keysight) U1252A DMM. This result is within 0.03% of the expected value, which shows that the meter's calibration is still good, and demonstrates the accuracy of the Multimeter Checker and Calibrator.

Summary

While we set out to add an AC voltage and frequency reference to an otherwise straightforward DC reference design, we think that being able to control the operation of the AC source manually will be a handy feature that many people will use. The USB interface also gives this handy little device a range of possible uses.

One thing to watch out for is noisy USB charger power supplies; they can cause frequency measurements of the ACV output to be unstable. In that case, the best solution is to power it from a USB power bank. A laptop USB port usually provides enough clean power to get stable readings from the *Multimeter Checker*.

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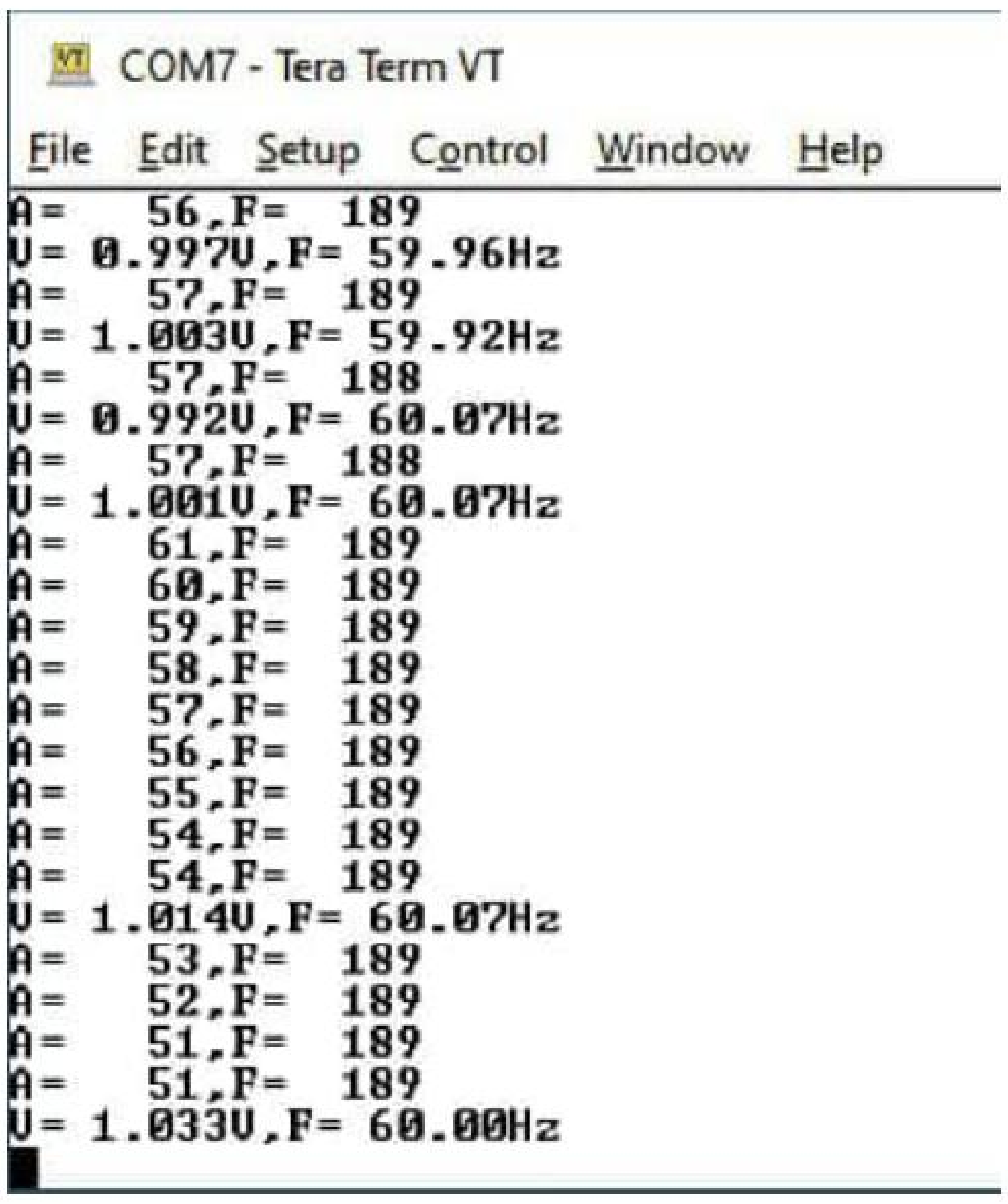

```
A=   56,F=  189
V= 0.997V,F= 59.96Hz
A=   57,F=  189
V= 1.003V,F= 59.92Hz
A=   57,F=  188
V= 0.992V,F= 60.07Hz
A=   57,F=  188
V= 1.001V,F= 60.07Hz
A=   61,F=  189
A=   60,F=  189
A=   59,F=  189
A=   58,F=  189
A=   57,F=  189
A=   56,F=  189
A=   55,F=  189
A=   54,F=  189
A=   54,F=  189
V= 1.014V,F= 60.07Hz
A=   53,F=  189
A=   52,F=  189
A=   51,F=  189
A=   51,F=  189
V= 1.033V,F= 60.00Hz
```

Screen 1: a typical output from the USB serial port. You can trigger the two-line reports shown here by pressing the space bar, while the single-line entries are due to manual changes in the amplitude and frequency settings. Mode changes do not produce any output but will be seen in changes to the illuminated LEDs on the Multimeter Checker.

Using Cheap Asian Electronic Modules

By Jim Rowe

MOS metal oxide semiconductor Air Quality Sensors

In the last few years we've noticed the rise of low-cost air quality sensors and sensing modules. In this article we'll investigate some of the currently available MOS (metal oxide semiconductor) type sensor modules, what they offer and how to use/interface them with an Arduino Uno.

MOS type air quality sensors (sometimes called 'MOx sensors') rely on the behaviour of particles of a metal oxide (usually tin oxide) when heated in the presence of air and/or other gases. The basic principle is shown in Fig.1, which depicts a cross-section of a typical MOS sensor.

The silicon substrate of the sensing chip has a thin layer of tin oxide on the top, placed there by chemical vapour deposition. Electrodes at each end allow its resistance to be measured. On the underside of the chip is a heater element, used to heat the oxide layer to around 200-250°C, to speed up the sensor's response.

When the oxide layer is heated in the presence of clean air, donor electrons in the oxide attract oxygen molecules from the air, and they are 'captured' by the oxide particles. As a result, a depletion layer forms on the surface of the oxide layer, and its electrical resistance rises.

But if reducing gases such as carbon monoxide (CO) and some volatile organic chemicals (VOCs) are present in the air, oxygen molecules in the surface of the oxide are released, and the depletion layer becomes thinner. As a result, the effective resistance of the oxide layer is reduced.

So the current passed by the oxide layer varies proportionally with the amount of reducing gas in the air surrounding the oxide layer. The higher the reducing gas level, the higher the current. Therefore, the basic MOS sensor essentially behaves as a reducing-gas-to-DC-analogue-current transducer.

The Hanwei MQ-135

Probably the most common of the low-cost MOS sensors currently available is the Hanwei MQ-135, which is designed to be sensitive to ammonia (NH_3), nitrous oxides (NO_x), carbon dioxide (CO_2), alcohol, benzene and smoke. Like the other sensors in the Hanwei series, the MQ-135 sensor comes in a cylindrical 6-pin package, 19mm in diameter and 15mm high.

Most modules using the MQ-135 simply take the current output from the sensor and convert it to a proportional voltage using a fixed load resistor. The output voltage can then be measured using a DMM, or fed into one of the ADC inputs of a microcontroller unit (MCU).

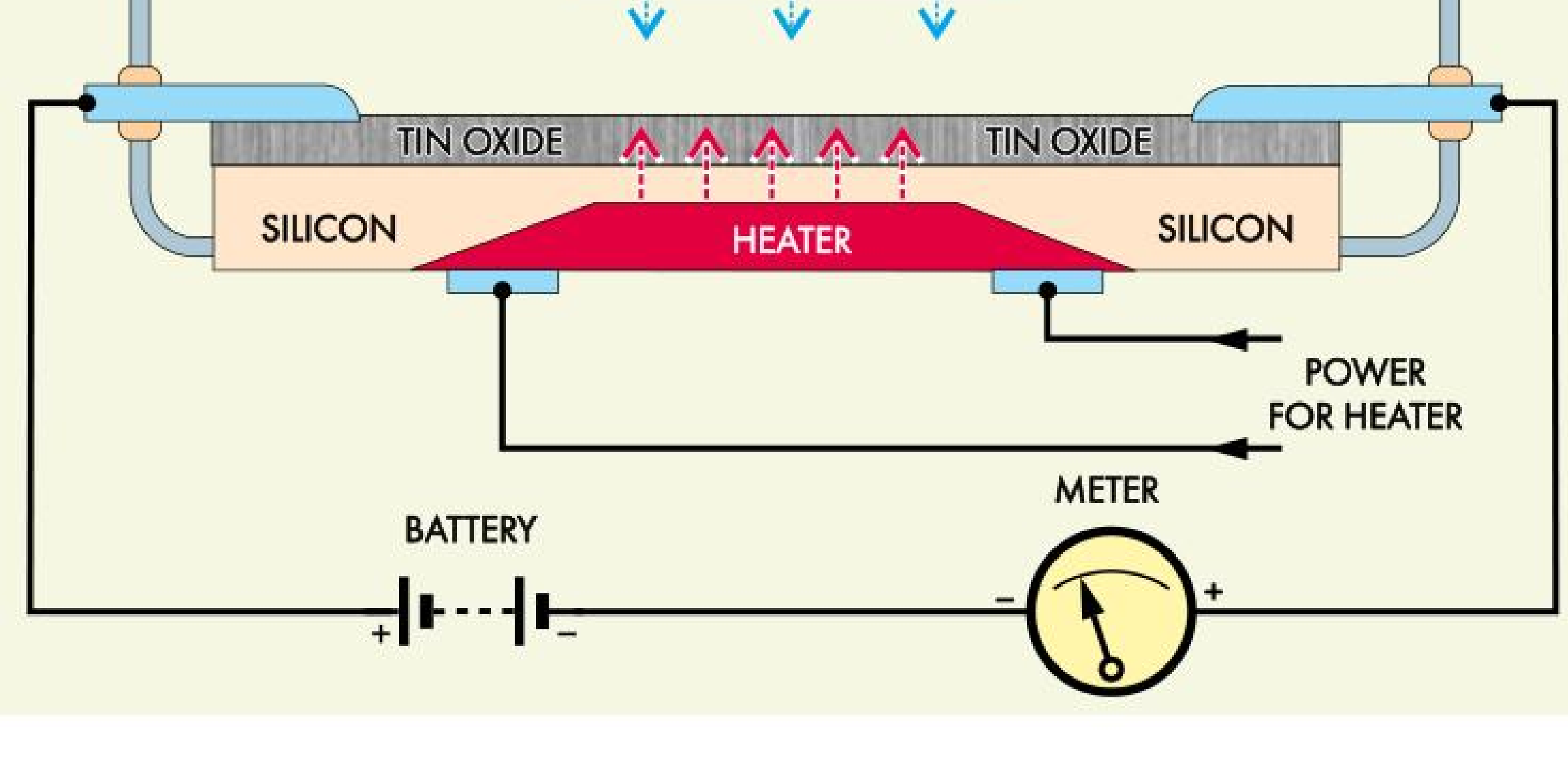

Fig.1: shows the cross-section of a MOS (metal oxide semiconductor) sensor and how it works.

Fig.2 shows the circuit of Hanwei's own air sensor module using the MQ-135. The MQ-135's heater pins (H) are connected between the +5V (V_{CC}) line and the GND line via a 5.1Ω series current-limiting resistor.

One end of the tin oxide sensing resistor (R_S) is connected to the +5V line via the two A pins, and the other end goes to the GND line via the two B pins and a 1kΩ load resistor. The two B pins are also connected to the A0 analogue output pin, to allow the voltage across the load resistor to be sent to a DMM or an MCU's ADC input.

The rest of the components are so that the module can also be used as a simple gas level alarm. One half of the LM393 dual comparator (IC1b) compares the voltage across the 1kΩ load resistor with a reference voltage set using trimpot VR1, so whenever the A0 voltage rises slightly above the reference voltage, the output of IC1b

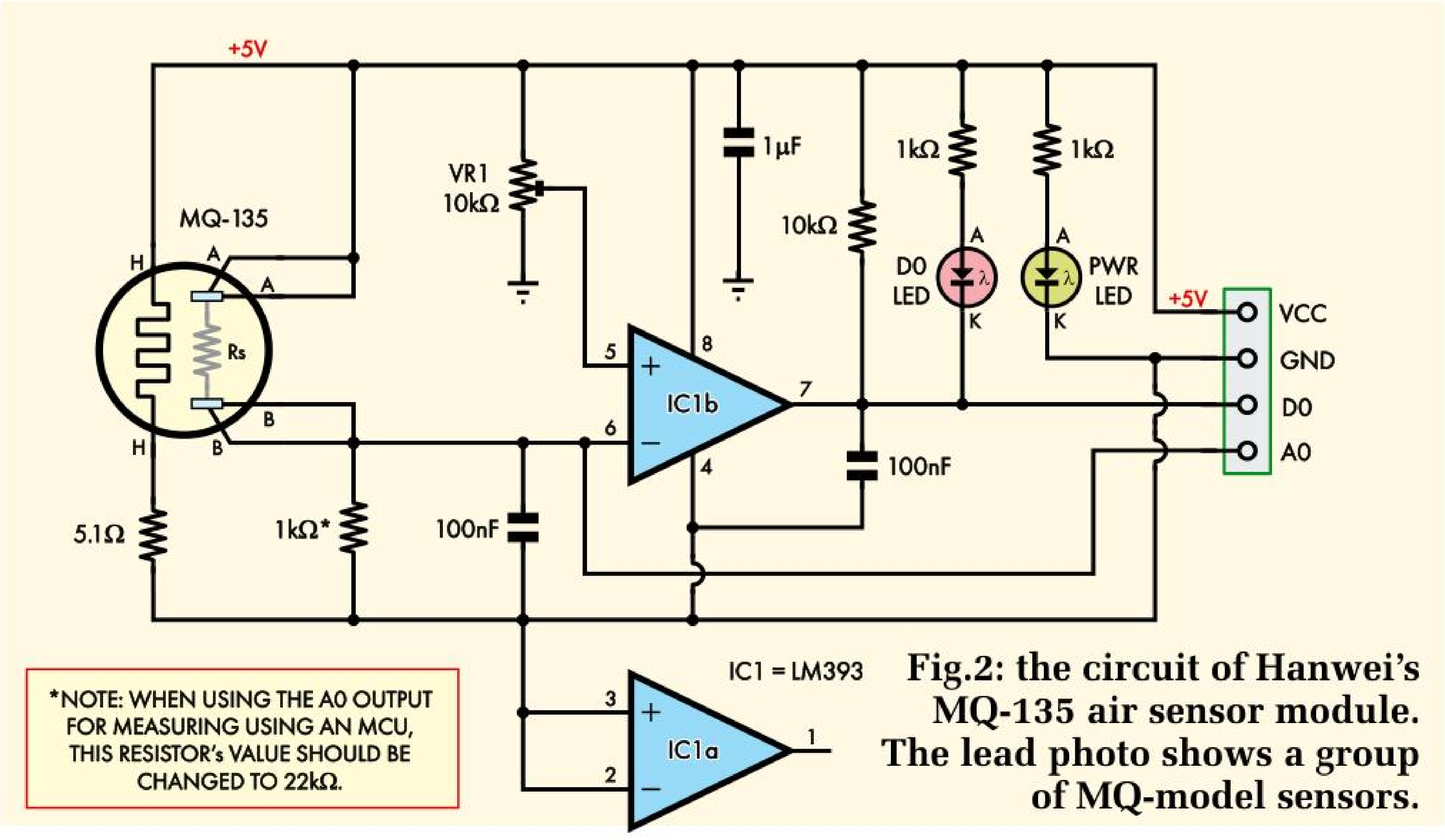
+5V
MQ-135
H
A
Rs
B
H
B
A
B
VR1
10kΩ
1µF
10kΩ
1kΩ
1kΩ
A
A
D0
LED
PWR
LED
K
K
+5V
VCC
GND
D0
A0
5
6
8
4
7
IC1b
100nF
5.1Ω
1kΩ*
100nF
IC1 = LM393
3
2
1
IC1a
*NOTE: WHEN USING THE A0 OUTPUT FOR MEASURING USING AN MCU, THIS RESISTOR's VALUE SHOULD BE CHANGED TO 22kΩ.

Fig.2: the circuit of Hanwei's MQ-135 air sensor module. The lead photo shows a group of MQ-model sensors.

(pin 7) drops to near ground level, causing the D0 LED to begin glowing.

The voltage level at the D0 output pin is pulled down simultaneously.

One change should ideally be made to the module if you want to use it with an MCU for monitoring the gas level, rather than simply using it as a gas level alarm. This involves replacing the sensor's 1kΩ load resistor with a 22kΩ resistor to give a higher output voltage swing and improve reading accuracy.

This resistor is an M2012/0805-size (2.0 x 1.2mm) SMD component, so you'll need a fine-tipped soldering iron and either a magnifying glass or a microscope.

Fig.3 shows how to hook it up to an Arduino Uno or a compatible MCU after making that change. You just need to connect the module's V_{CC} and GND pins to the corresponding pins on the Arduino, plus the module's A0 pin to one of the Arduino's ADC input pins; in this case, we're using A2.

MQ-135 Sketch Program

```
void setup() {
 Serial.begin(9600);
 Serial.println(“Practical
  Electronics MQ-135
  demo!”);
}

void loop() {
 int reading =
  analogRead(A2);
 Serial.println(reading);
 delay(1000);
}
```

Sample Output

```
Practical Electronics MQ-135
  demo!
696
694
694
691
692
710
```

There are quite a few Arduino libraries and sketches available to work with the MQ-135 module. You'll find links to some of them in the list of links at the end of this article. However, I found many of them a bit tricky to negotiate.

But I did find some very helpful information on Rob's blog at: **https://blog.robberg.net/mq-135-arduino** Then I came across an elementary sketch using no libraries, but just showing the current analogue voltage provided at the module's A0 pin, see: **https://arduinolearning.com/amp/code**

I adapted this sketch slightly, and its listing is replicated below along with some of the sample output from when this sketch is running. When I breathed on the MQ-135, that caused the voltage reading to rise from under 700 to about 728 before falling back down again.

As you can see, there's no attempt to convert the A0 voltage readings to equivalent gas levels – for that, you would need one of the fancier sketches relying on their dedicated libraries.

SGX Sensortech MiCS-5524

Another MOS sensor found in low-cost air/gas sensing modules is the MiCS-5524, made by SGX Sensortech (an Amphenol company) in Switzerland. This is much smaller than the MQ-135, coming in an SMD package measuring only 7 x 5 x 1.6mm.

The MiCS-5524 detects CO, ethanol, hydrogen, ammonia and methane. It is used in an 18 x 13mm gas sensing module with the same name available from various internet suppliers, including ALiExpress, Mouser and PiHut for around £10 to £15).

The circuit of the MiCS-5524 module is shown in Fig.4. It is essentially just the sensor itself with an 82Ω current-limiting resistor for the sensor's heater and a 91kΩ load resistor for its sensing resistor R_S, with a 100nF capacitor across the latter to provide noise reduction.

P-channel MOSFET Q1 is included so that the power to the sensor can be controlled using the module's EN pin. This pin can be left floating if the module is to operate continuously.

Fig.5 shows how easy it is to connect the MiCS-5524 module to an Arduino Uno, while the sketch is shown overleaf with the sample output. The sketch is almost identical to the MQ-135 program and is similarly based on: **https://arduinolearning.com/amp/code**

SGX Sensortech MiCS-VZ-89TE

SGX Sensortech also makes a fancier and slightly larger module (23 x 14mm) called the MiCS-VZ-89TE, available from suppliers like element14, Digi-Key and Mouser for around £20 plus delivery. This module incorporates its own dedicated MCU with ADCs (analogue-to-digital converters) and embedded conversion algorithms.

As a result, this module can provide both PWM and I^{2}C digital outputs for CO_2 equivalent and TVOC (isobutylene equivalent).

I couldn't find any circuit diagram for the MiCS-VZ-89TE module, but its layout is shown in Fig.6.

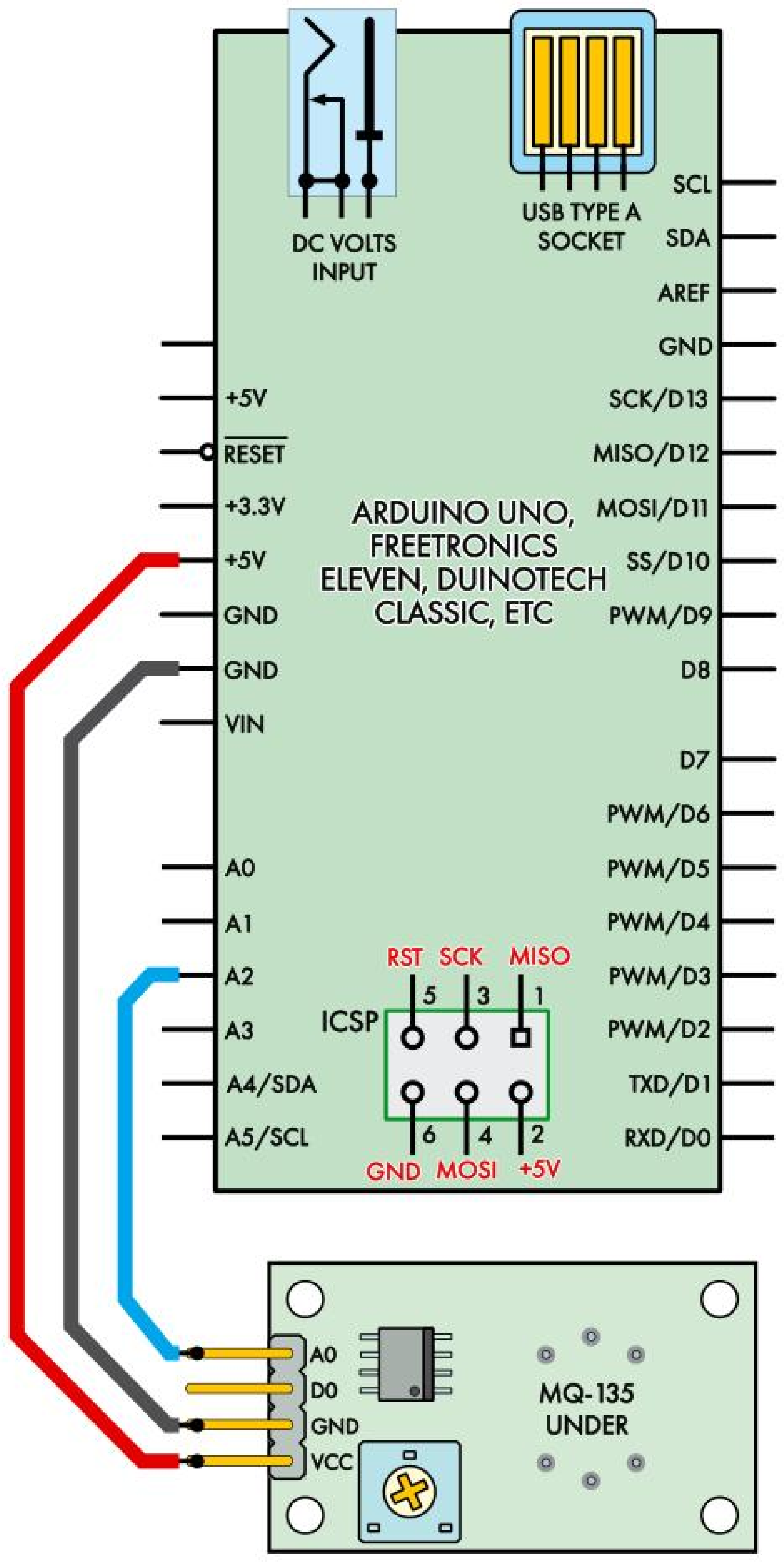
DC VOLTS
INPUT
USB TYPE A
SOCKET
SCL
SDA
AREF
GND
SCK/D13
MISO/D12
MOSI/D11
SS/D10
PWM/D9
D8
D7
PWM/D6
PWM/D5
PWM/D4
PWM/D3
PWM/D2
TXD/D1
RXD/D0
+5V
RESET
+3.3V
+5V
GND
GND
VIN
A0
A1
A2
A3
A4/SDA
A5/SCL
ARDUINO UNO,
FREETRONICS
ELEVEN, DUINOTECH
CLASSIC, ETC
ICSP
RST
SCK
MISO
5
3
1
6
4
2
GND
MOSI
+5V
A0
D0
GND
VCC
MQ-135
UNDER

Fig.3: the connection diagram for the MQ-135 sensor module with an Arduino Uno or similar.

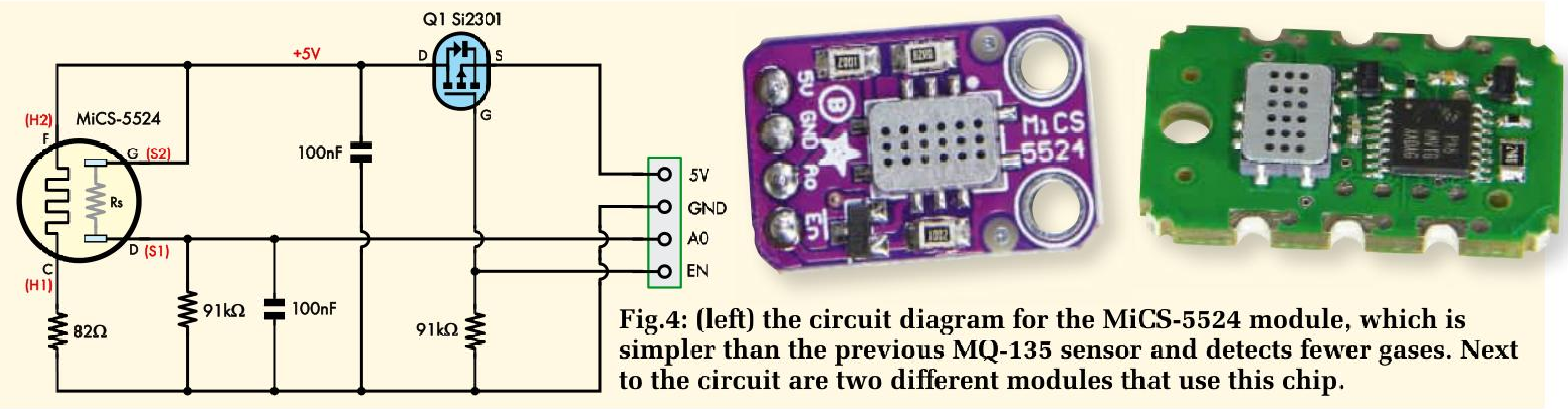
Q1 Si2301
+5V
D
S
G
(H2)
F
MiCS-5524
G (S2)
Rs
D (S1)
C
(H1)
100nF
5V
GND
A0
EN
82Ω
91kΩ
100nF
91kΩ
MiCS
5524

Fig.4: (left) the circuit diagram for the MiCS-5524 module, which is simpler than the previous MQ-135 sensor and detects fewer gases. Next to the circuit are two different modules that use this chip.

I found it fairly easy to connect to this module. I used two 5-pin sections of SIL header strip, with the top of the second and fourth pins of each strip cut short, allowing the tops of the remaining three pins to be soldered to the notches on one side of the module. You can then plug the complete assembly into a small breadboard for testing and use.

Fig.7 shows how the MiCS-VZ-89TE module can be connected to an Arduino Uno or equivalent MCU. The GND connection goes to one of the Arduino's GND pins, while the module's 3.3V power connection goes to the Arduino's +3.3V pin.

The module's I^{2}C connections SDA and SCL are wired to the Arduino's pins A4/SDA and A5/SCL, respectively. Each of these pins needs an external 4.7kΩ pull-up resistor connected to the +3.3V pin, because the MiCS-VZ-89TE module doesn't provide pull-ups.

I found an Arduino sketch and library to read the CO$_2$ and VOC levels from a MiCS-VZ-89TE, written by H Grabas – it is available on his website at: **https://bit.ly/pe-jul23-grab**

This sketch and its library worked so well that I adapted it to produce the sketch listed opposite, along with a sample of the output from the Arduino IDE Serial Monitor. For this to work, you need to download Mr Grabas' library from his website and install it as a library in the Arduino IDE.

When running, it gives you a VOC reading and a CO$_2$ reading approximately once per second. I tried breathing on the module's sensor, causing the VOC readings to rise to around 270.4ppb (parts per billion), while the CO$_2$ reading barely moved from about 414 ppm (parts per million).

Next, I sprayed a tiny amount of isopropanol (spectacle cleaning fluid) a few centimetres above the sensor, causing the VOC reading to jump up to its maximum figure of 1000ppb. So the MiCS-VZ-89TE and the sketch and library certainly seem to be working.

The ScioSense CCS811

Another MOS sensor found in several low-cost air/gas sensing modules is the CCS811, made by ScioSense BV in Eindhoven, The Netherlands. The CCS811 is in a tiny SMD package, measuring just 4 x 3 x 1.2mm.

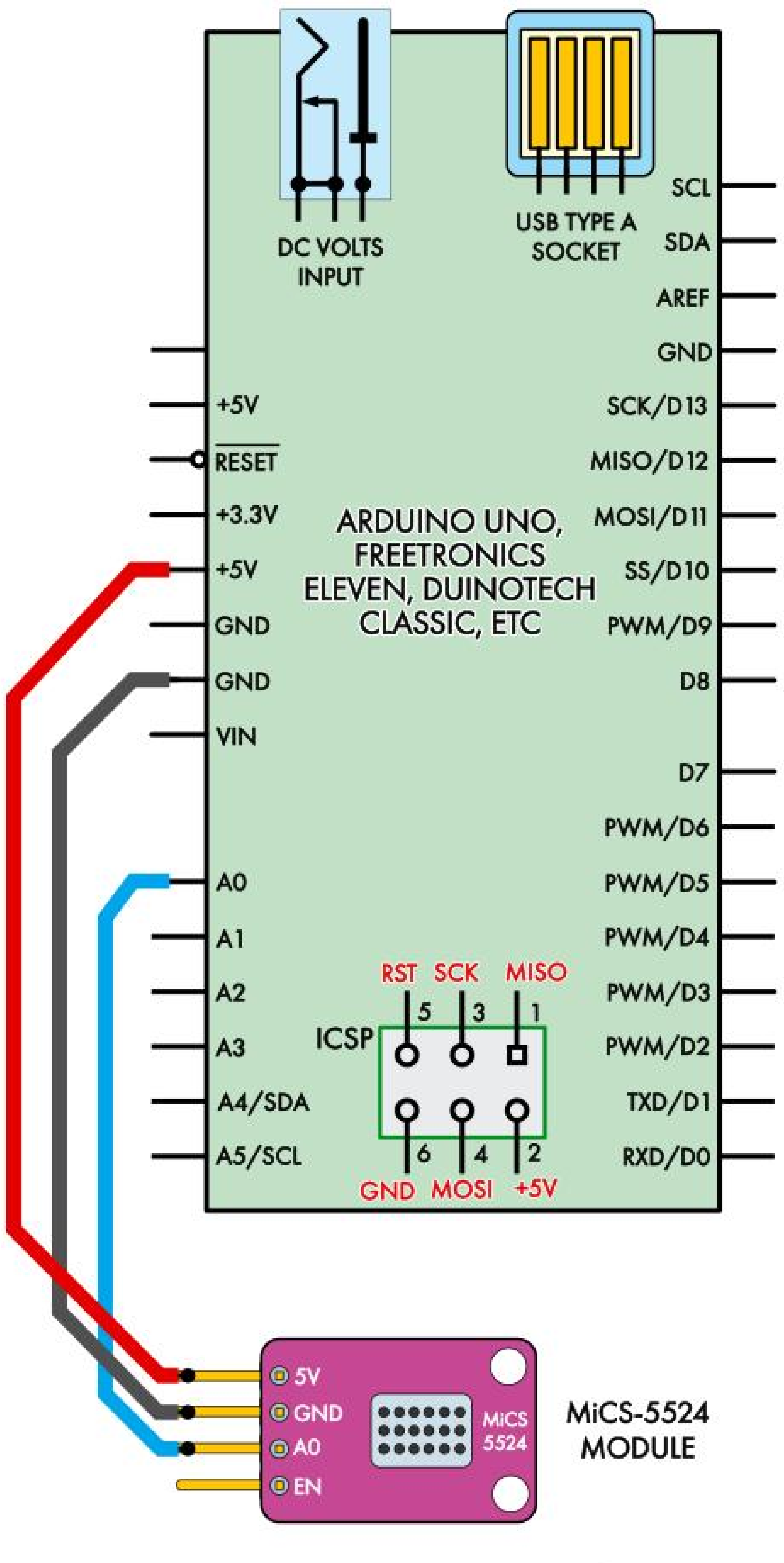
DC VOLTS INPUT
USB TYPE A SOCKET
SCL
SDA
AREF
GND
SCK/D13
MISO/D12
MOSI/D11
SS/D10
PWM/D9
D8
D7
PWM/D6
PWM/D5
PWM/D4
PWM/D3
PWM/D2
TXD/D1
RXD/D0
+5V
RESET
+3.3V
+5V
GND
GND
VIN
A0
A1
A2
A3
A4/SDA
A5/SCL
ARDUINO UNO, FREETRONICS ELEVEN, DUINOTECH CLASSIC, ETC
ICSP
RST SCK MISO
5 3 1
6 4 2
GND MOSI +5V
5V
GND
A0
EN
MiCS 5524
MiCS-5524 MODULE

Fig.5: MiCS-5524 connection diagram.

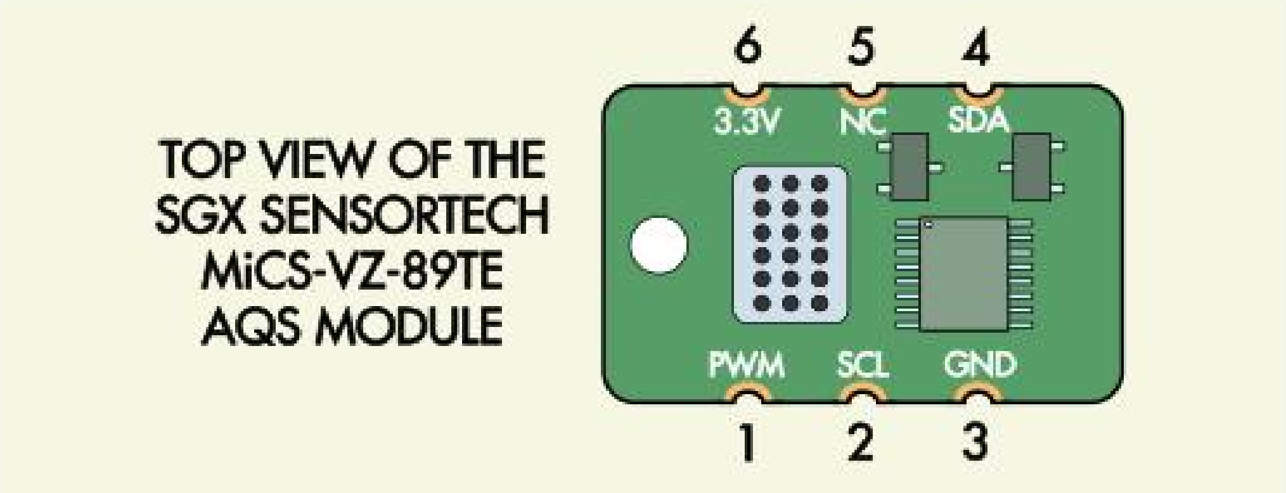
TOP VIEW OF THE SGX SENSORTECH MiCS-VZ-89TE AQS MODULE
6 5 4
3.3V NC SDA
PWM SCL GND
1 2 3

Fig.6: the layout diagram for the MiCS-VZ-89TE module.

MiCS-5524 Sketch

```
void setup() {
 Serial.begin(9600);
 Serial.println(“Practical
  Electronics MiCs-5524
  demo!”);
}

void loop() {
 int reading =
  analogRead(A0);
 Serial.println(reading);
 delay(1000);
}
```

Sample Output

```
Practical Electronics MiCs-
  5524 demo!
40
39
40
39
```

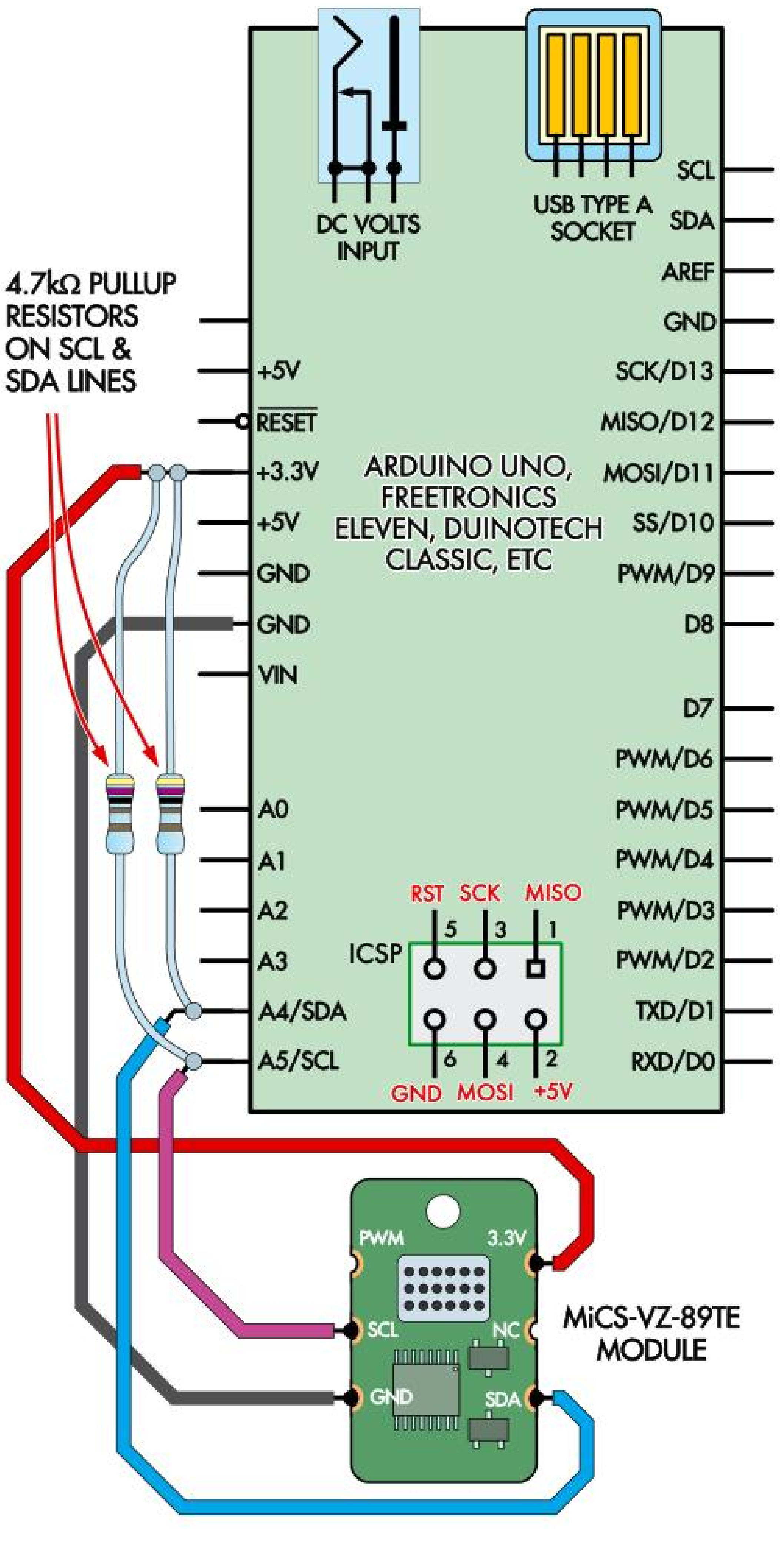
DC VOLTS INPUT
USB TYPE A SOCKET
4.7kΩ PULLUP RESISTORS ON SCL & SDA LINES
SCL
SDA
AREF
GND
SCK/D13
MISO/D12
MOSI/D11
SS/D10
PWM/D9
D8
D7
PWM/D6
PWM/D5
PWM/D4
PWM/D3
PWM/D2
TXD/D1
RXD/D0
+5V
RESET
+3.3V
+5V
GND
GND
VIN
A0
A1
A2
A3
A4/SDA
A5/SCL
ARDUINO UNO, FREETRONICS ELEVEN, DUINOTECH CLASSIC, ETC
ICSP
RST SCK MISO
5 3 1
6 4 2
GND MOSI +5V
PWM
3.3V
SCL
NC
GND
SDA
MiCS-VZ-89TE MODULE

Fig.7: MiCS-VZ-89TE connection diagram to an Arduino Uno.

MiCS-VZ-89TE Sketch:

```
#include <MICS-VZ-89TE.h>
#include <Wire.h>

MICS_VZ_89TE voc;

void setup() {
 voc.begin();
 Serial.begin(9600);
 Serial.println(“Reading the
  MiCS-VZ-89TE sensor”);
}

void loop() {
 voc.readSensor();
 Serial.print(“VOC =”);
 Serial.print(voc.getVOC());
 Serial.print(“ | ”);
 Serial.print(“CO2 = ”);
 Serial.println(voc.
  getCO2());
 delay(1000);
}
```

Sample Output

```
Reading the MiCS-VZ-89TE
  sensor
VOC = 0.00    | CO2 = 413.97
VOC = 135.37  | CO2 = 413.97
VOC = 270.74  | CO2 = 413.97
VOC = 1000.00 | CO2 = 420.96
```

Despite its miniscule size, this sensor incorporates both an ADC and a dedicated MCU with built-in conversion algorithms, plus an I²C digital interface to link directly to a PC or an MCU like an Arduino or a Micromite. It's described by ScioSense as an 'ultra-low power digital gas sensor' and detects a range of VOCs and provide both eTVOC (equivalent total VOC) and eCO$_2$ (equivalent CO$_2$) levels.

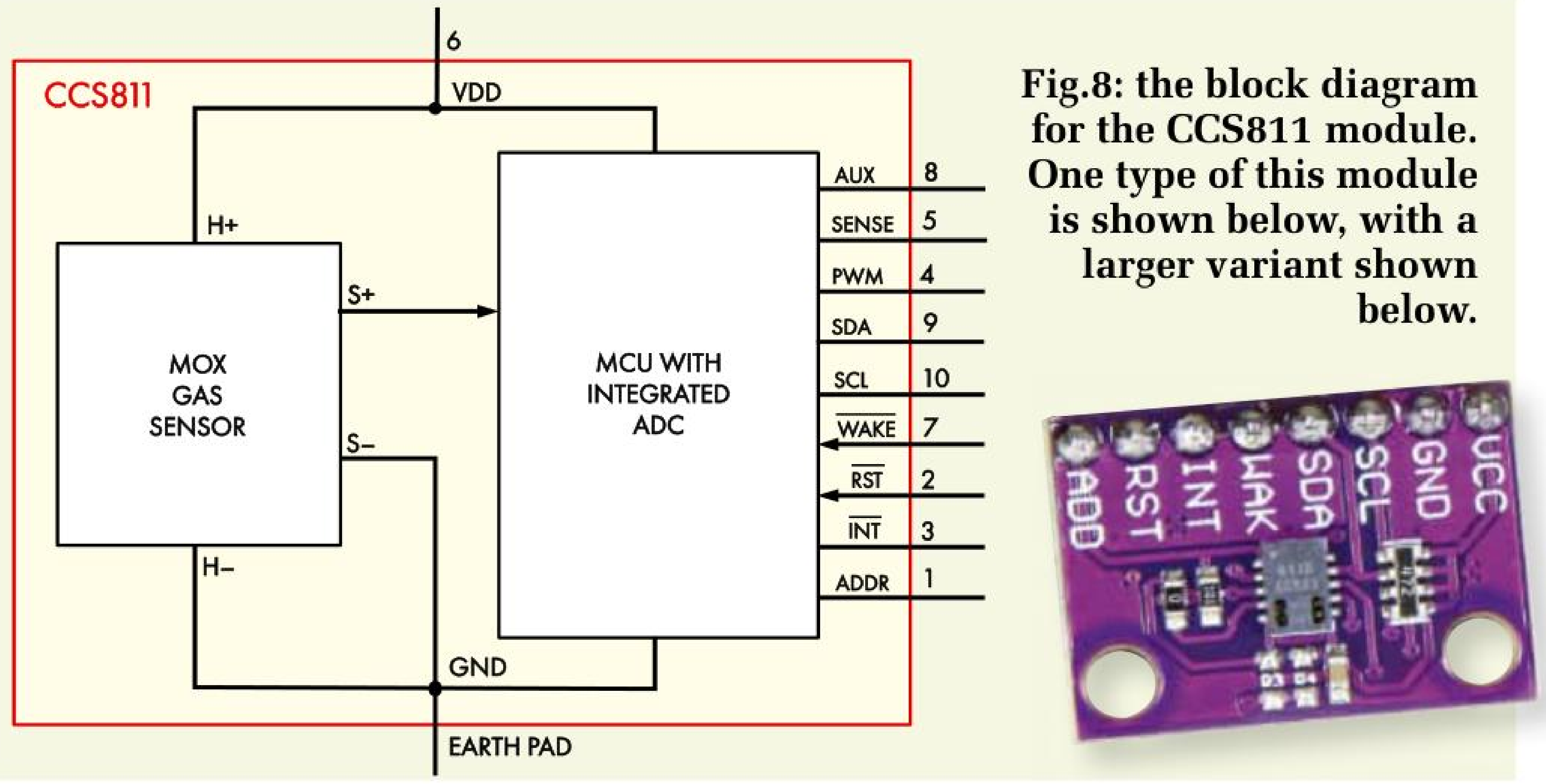
CCS811
6
VDD
H+
S+
S−
H−
MOX GAS SENSOR
MCU WITH INTEGRATED ADC
AUX 8
SENSE 5
PWM 4
SDA 9
SCL 10
WAKE 7
RST 2
INT 3
ADDR 1
GND
EARTH PAD
ADD
RST
INT
WAK
SDA
SCL
GND
VCC

Fig.8: the block diagram for the CCS811 module. One type of this module is shown below, with a larger variant shown below.

Fig.8 is a block diagram of the CCS811. Pin 4 (PWM) and pin 5 (SENSE) must be connected together for correct operation of the MOX sensor's heater control circuit. Pin 1 (ADDR) is to allow the CCS811's I²C address to be set to either 90d/5Ah (ADDR pin low) or 91d/5Bh (ADDR pin high), while the AUX pin (8) has no internal connection.

The CCS811 sensor is used in many air quality sensing modules, including the Keyestudio KS0457 CO$_2$ Air Quality module, the Duinotech SEN-CCS811 Air Quality Sensor module (Jaycar Cat XC3782), the Adafruit CCS811 Air Quality Sensor and the CJMCU-811 CO$_2$, Temperature and Humidity Sensor from Banggood.

Fig.9 shows the circuit for many of these CCS811 sensor modules. Along with the CCS811 sensor itself, there's voltage regulator REG1, which steps down the incoming +5V power to provide the 3.3V needed by the CCS811, plus MOSFETs Q1 and Q2 which, together with four 10kΩ pull-up resistors, perform logic-level conversion for the I²C digital communication lines (SDA and SCL).

Diodes D1 and D2, together with the two 100kΩ pull-up resistors, allow the WAKE and RST pins of the CCS811 to be pulled low. The WAKE pin must be pulled to ground to allow the chip to operate.

Note that pin 1 of the CCS811 is pulled low by a 100kΩ resistor to set the I²C address to 90d/5Ah. Also, as mentioned earlier, pins 4 and 5 are tied together and pulled high via two more 100kΩ resistors.

Incidentally, some CCS811-based modules (such as the CJMCU-811) have an additional pin on the I/O connector, with the extra pin connected to pin 1 of the CCS811 and labelled 'ADD'. This allows the I²C address of the module to be changed to 91d/5Bh by pulling the pin high.

It's quite easy to connect the SEN-CCS811 and most of the other CCS811-based air quality modules to an MCU like an Arduino Uno, as shown in Fig.10. The V$_{CC}$ and GND pins connect to the +5V and GND pins of the

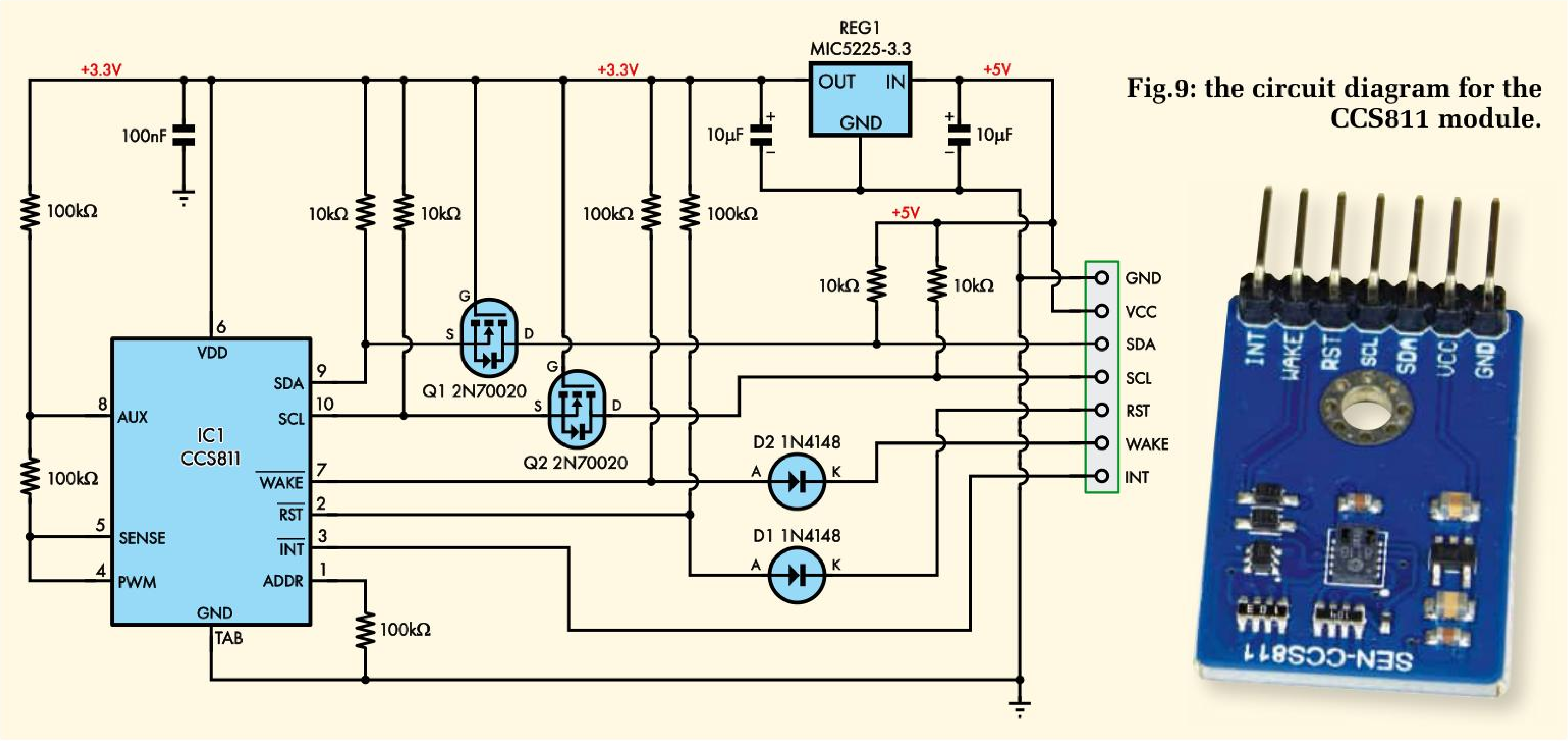
REG1
MIC5225-3.3
OUT IN
GND
+3.3V
+3.3V
+5V
+5V
100nF
10µF
10µF
100kΩ
100kΩ
10kΩ
10kΩ
100kΩ
100kΩ
10kΩ
10kΩ
100kΩ
IC1
CCS811
VDD
AUX
SENSE
PWM
GND
TAB
SDA
SCL
WAKE
RST
INT
ADDR
Q1 2N70020
Q2 2N70020
G
S
D
D2 1N4148
D1 1N4148
A
K
GND
VCC
SDA
SCL
RST
WAKE
INT
SEN-CCS811

Fig.9: the circuit diagram for the CCS811 module.

PoLabs

For more information or software download please visit

www.poscope.com/epe

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Arduino, while the SDA pin goes to the Arduino's A4/SDA pin and the SCL pin to the Arduino's A5/SCL pin. Finally, the module's WAKE pin connects to another GND pin on the Arduino.

Some modules have the pins in a different order, so make sure you check the connections for the module you are using.

Several Arduino libraries are available to support a sketch communicating with these modules. I found the easiest one to use was the Keyestudio KS0457 library (**CCS811.h** and **CCS811.cpp**), available from: **https://fs.keyestudio.com/KS0457**

I also downloaded Keyestudio's **readData.ino** sketch and adapted it to produce my own version of the sketch: **read_CCS811_data.ino**. It's a bit too long to reproduce the listing here, but you can download it from the July 2023 page of the *PE* website at: **https://bit.ly/pe-downloads**

Shown below is the output of that sketch. The Arduino provides a stream of measurements for both the eCO_2 level in ppm and eVOC in ppb.

At one point, I blew in the direction of the CCS811 sensor to give it some extra CO_2. That's the reason for the sudden rise in eCO_2 and eTVOC readings, from around 400ppm and 1-2ppb up to peaks of 1743ppm and 384ppb a second later. Then the readings fell slowly after that.

Summary

After trying several of these modules, I'm less keen on those based on the MQ-135 sensor than on the Sensortech MiCS sensors or the ScioSense CCS811 sensor. That's mainly because of the scarcity of easy-to-understand software if you want to do more than simply 'raise the alarm' if the CO_2/VOC level rises above a preset 'safe' level.

I'm also not that keen on modules based on the SGX Sensortech MiCS-5524 sensor for much the same reason.

Overall, I prefer the 'smarter' modules like the SGX Sensortech MiCS-VZ-89TE or most of those using the CCS811 sensor. These modules are all much easier to get going with an MCU like an Arduino as a reliable CO_2/VOC sensor.

Sample Output

```
Getting data from the
  CCS811...
eCO2: 400ppm, eTVOC: 0ppb
eCO2: 410ppm, eTVOC: 1ppb
eCO2: 414ppm, eTVOC: 2ppb
eCO2: 1743ppm, eTVOC: 384ppb
eCO2: 1345ppm, eTVOC: 143ppb
eCO2: 977ppm, eTVOC: 87ppb
```

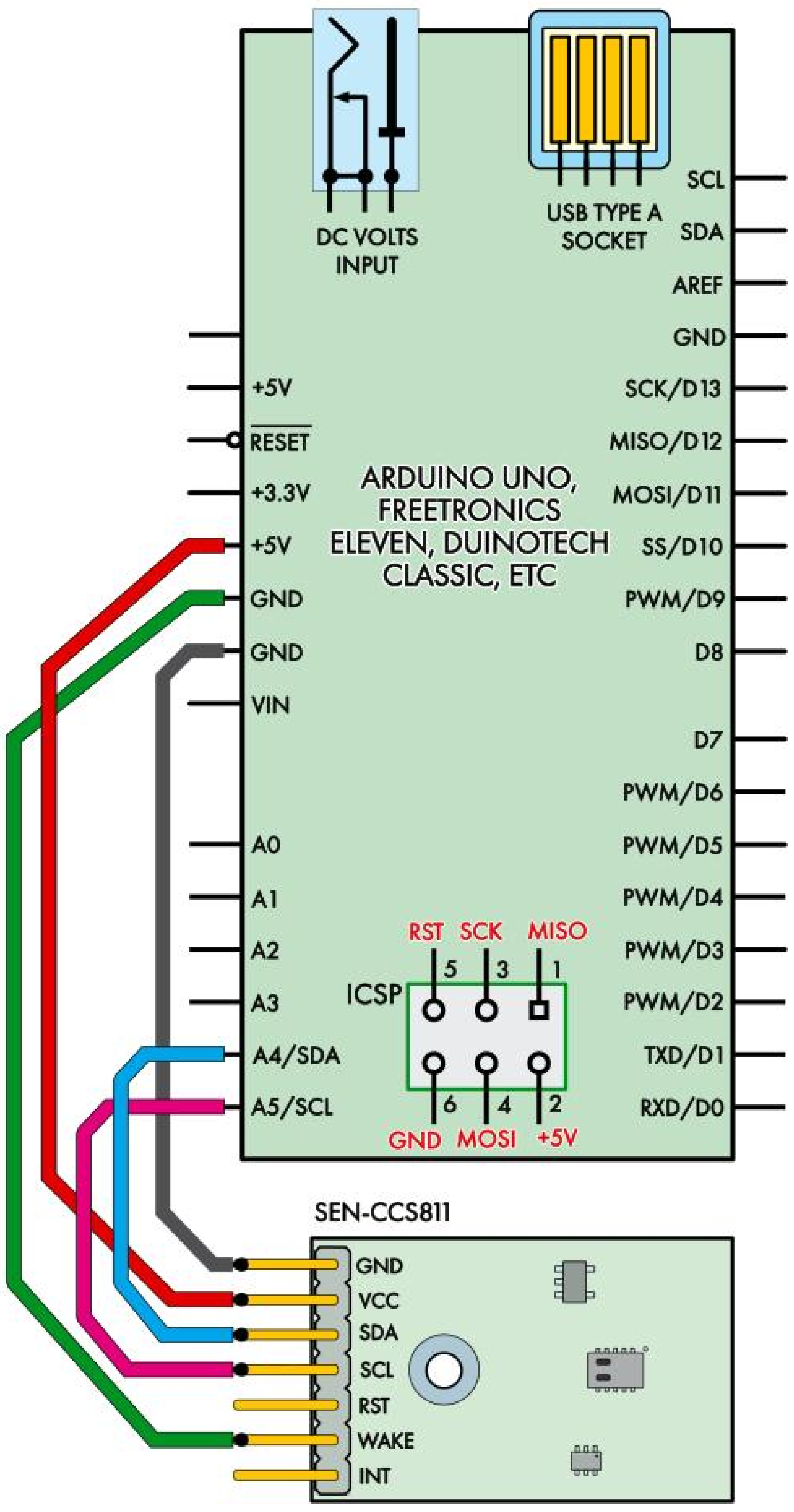

Fig.10: CCS811 connection diagram.

I would give first prize to the MiCS-VZ-89TE module (element14 2925865). But second prize would go to any of the modules based on the ScioSense CCS811 sensor, like the Duinotech SEN-CCS811 from Jaycar (Cat XC3782), the CJMCU-811 from Banggood (ID 1157216), the Keyestudio KS0457 or the Adafruit CCS811: **www.adafruit.com/product/3566**

I will describe some of the NDIR and PAS type air quality sensor modules in a future article.

Useful Links

MQ-135:
- **www.arduinolibraries.info/libraries/mq135**
- **https://github.com/Phoenix1747/MQ-135**
- **siliconchip.com.au/link/abct**
- **https://blog.robberg.net/mq-135-arduino**

MiCS-5524:
- **www.sgxsensortech.com**
- **siliconchip.com.au/link/abcu**
- **https://bit.ly/pe-jul23-grab**

CCS811:
- **https://fs.keyestudio.com/KS0457**
- **https://bit.ly/pe-jul23-811**

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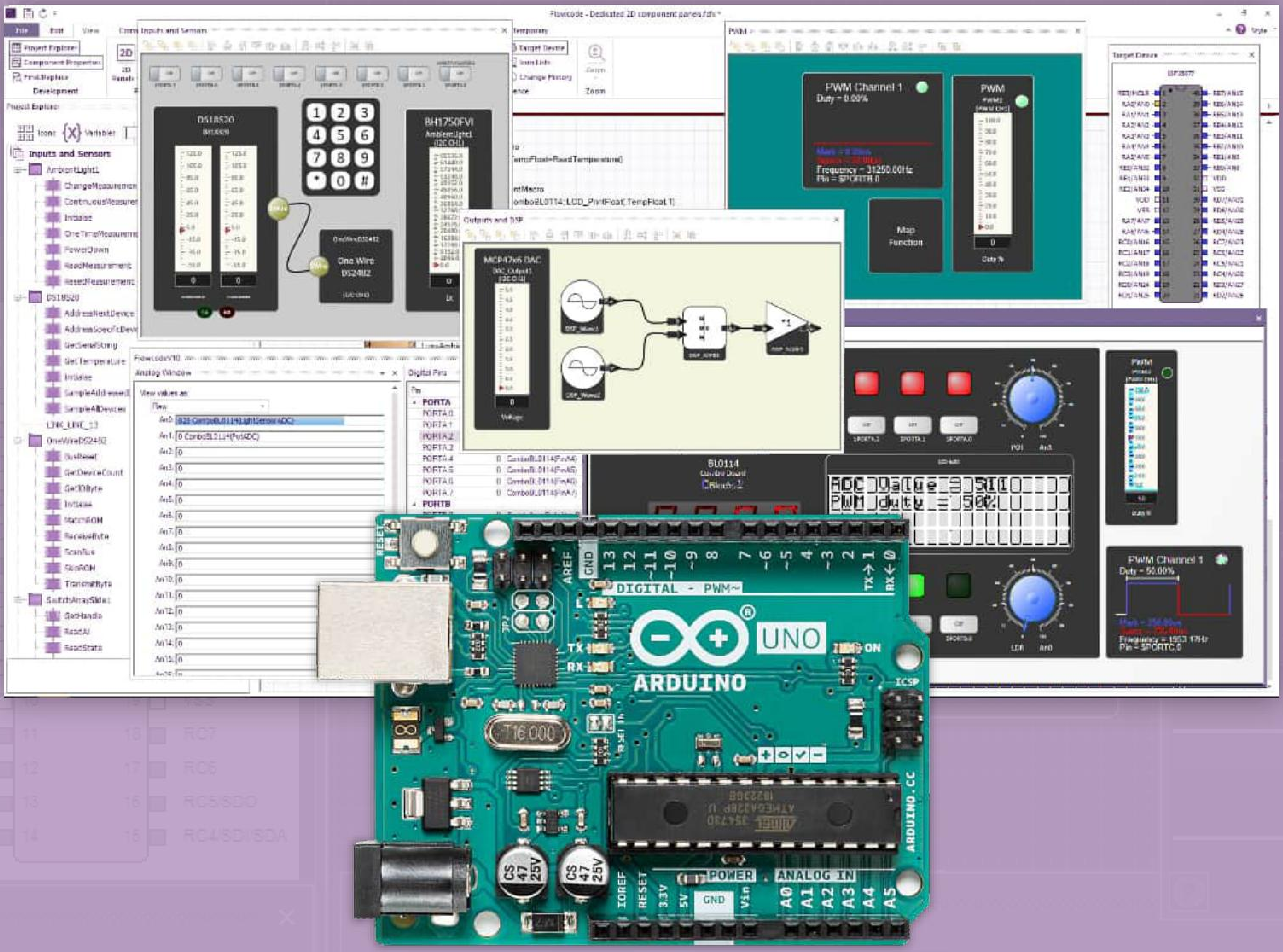

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KickStart

by Mike Tooley

Part 12: Your first 'scope

Our occasional *KickStart* series aims to show readers how to use readily available low-cost components and devices to solve a wide range of common problems in the shortest possible time. Each of the examples and projects can be completed in no more than a couple of hours using 'off-the-shelf' parts. As well as briefly explaining the underlying principles and technology used, the series will provide you with a variety of representative solutions and examples along with just enough information to be able to adapt and extend them for their own use.

This twelfth instalment deviates from earlier parts by taking time out to introduce you to an item of test equipment that will undoubtedly become invaluable as you progress with your interest in electronics. We provide some valuable pointers for those who are about to purchase an oscilloscope for the first time or who might be planning to upgrade an instrument in the near future.

I have very fond memories of my first oscilloscope. Built on the kitchen table more than 60 years ago, this rather dangerous instrument used a 3.5-inch (90mm) ex-surplus VCR138 cathode ray tube (CRT) together with an assortment of octal-based valves and a 1.2kV high-tension power supply. Despite the challenge in assembling it, the instrument proved to be something of a revelation, providing me with an insight into what was going on inside the circuits I was experimenting with. Not surprisingly, I quickly became convinced that a 'scope was going to be a permanent addition to my test bench. So, flushed with success, I set about designing an improved version using a 5-inch VCR97 tube with an even more dangerous 2kV supply!

Since then, I've worked my way through at least two dozen 'scopes of different types and specification, some costing as little as £25 to several costing well over £5,000. My current inventory of test equipment includes a fast digital storage oscilloscope (DSO), a general-purpose USB 'scope that connects with my desktop PC, laptop, tablet and phone, and three elderly CRT-based scopes that adorn the shelves in my workshop.

What can an oscilloscope do?

Before we describe the range of instruments that you might want to consider as candidates for a 'first 'scope', it's worth pondering on what the instrument can do, and what it might do for you. The key point is that an oscilloscope can provide a plethora of information about what is going on in a circuit. In effect, it will allow you to 'see' into the circuit, displaying waveforms that correspond to the signals and voltages present. Oscilloscopes can be used in a wide variety of situations, including:

- Display of signal waveforms: the shape of a wave can indicate the absence of linearity and the presence of distortion, as well as the presence of noise, unwanted harmonics, and other undesirable signal components
- Measurement of signal amplitude (peak or peak-to-peak values)
- Measurement of signal frequency/period
- Measurement of time delay and the rise and fall time of rectangular pulses
- Measurement of phase change and phase difference between signals
- Measurement of dc levels and offsets that may be superimposed on ac signals (it is possible to measure dc voltage using an oscilloscope, but an ordinary bench voltmeter will invariably do a much better job)
- Investigation of transients and other non-repetitive short-duration phenomena
- Observation of drift and other long-term variations that may affect signal characteristics such as amplitude, frequency and phase.

Choosing an instrument

For most electronics enthusiasts, the acquisition of an oscilloscope represents a significant long-term investment. It therefore merits careful consideration based on several

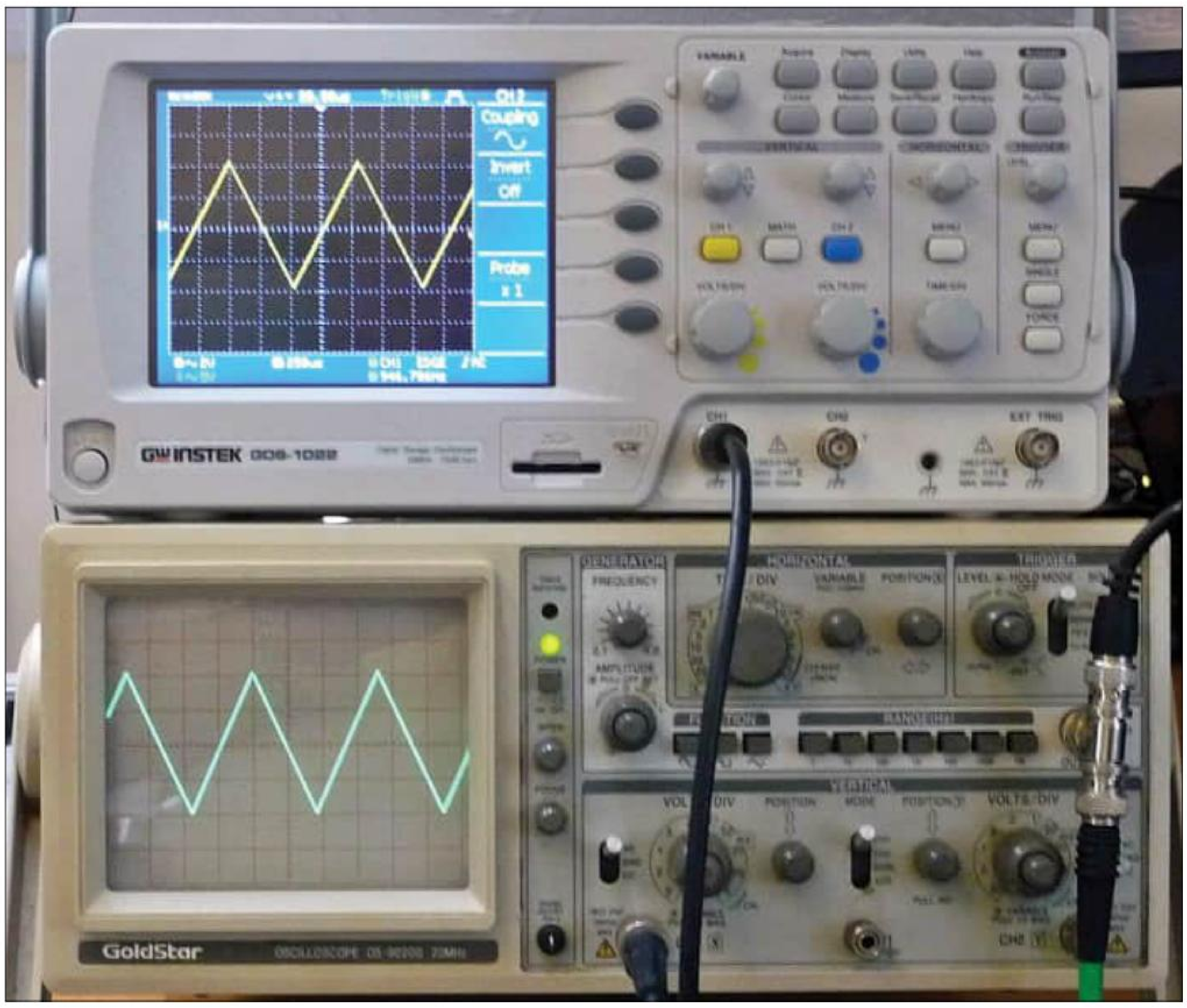


Fig.12.1. Conventional analogue (bottom) and digital storage 'scope (top) displaying the same waveform.

factors, paramount among these is understanding your current *and* future test and measurement requirements. Unfortunately, due to a wide variation in features, displays, controls and specifications offered, the choice of instrument can often be bewildering.

The main categories of oscilloscope can be summarised as follows:

- Self-contained digital storage 'scopes (DSO) in both bench and portable versions
- DSO with a USB connection to a host PC, laptop or tablet computer
- Sound card 'scope software running on PC, laptop or tablet computer
- Older, analogue CRT-based oscilloscopes.

Within each of these main categories there are a host of sub-categories. For example, 'scopes with multi-channel capability, high-resolution 'scopes, 'scopes with built-in signal/waveform generators, and 'scopes designed specifically for automotive and other specialised applications.

DSOs combine elements of both hardware and software that work together to provide all the functionality of a conventional 'scope combined with additional features such as spectrum analysis, data logging, and digital measurement of voltage, time and frequency. In many cases you may find that a DSO will augment or replace several items of existing test equipment. However, despite these plus points, it might be worth considering some of the alternatives before homing in on a bench DSO.

Available in plentiful quantities on the second-user market, an older CRT-based oscilloscope can be a very cost-effective solution. Whilst such instruments can deliver excellent performance at low cost, they are unlikely to have features such as waveform storage, on-screen measurement data and automatic adjustment of controls. One other consideration worth mentioning is the relatively large size and weight of CRT-based instruments when compared with modern DSO. If you can live with these limitations, you might find that a second-hand analogue instrument is a good and inexpensive choice.

Another alternative might be a USB DSO. These budget-priced instruments offer a compact alternative to bulky and relatively expensive benchtop instruments. You can now fit a 200MHz, 1 GS/s instrument together with probes and accessories into a laptop bag and new functionality can be delivered through manufacturer software upgrades and active community support.

A PC soundcard-based 'scope can be a viable alternative if you are only interested in displaying audio frequency waveforms over a limited frequency range (eg, 100Hz to 10kHz). A high-performance sound card with fast sampling and low noise will give better results but you will need a means of calibrating the 'scope (see *Getting started* on p.45). You should also note that, while soundcard 'scopes can work well for repetitive waveforms, triggering can be difficult when displaying irregular waveforms and pulses. Measurements of rise- and fall-times will almost certainly be wildly inaccurate. One further cautionary note is that the input impedance offered by a soundcard will typically be around 50kΩ. This is very much lower than the standard 1MΩ associated with conventional instruments.

Questions that you need to answer

To help you identify a suitable type of instrument it is worth answering a few questions.

1. Where will you use the instrument?

Will the 'scope only be used on the bench where an AC supply is available, or do you need a portable instrument that can operate from internal batteries?

2. Would you consider an older second-hand instrument as an alternative to a more modern digital instrument?

Second-hand instruments of good specification are readily available and can usually be obtained very cheaply. However, due to the use of a cathode ray tube (CRT) display, older second-hand instruments tend to be quite bulky, relatively heavy, and often require a significant amount of bench space. They may also be lacking several of the more useful features available from digital

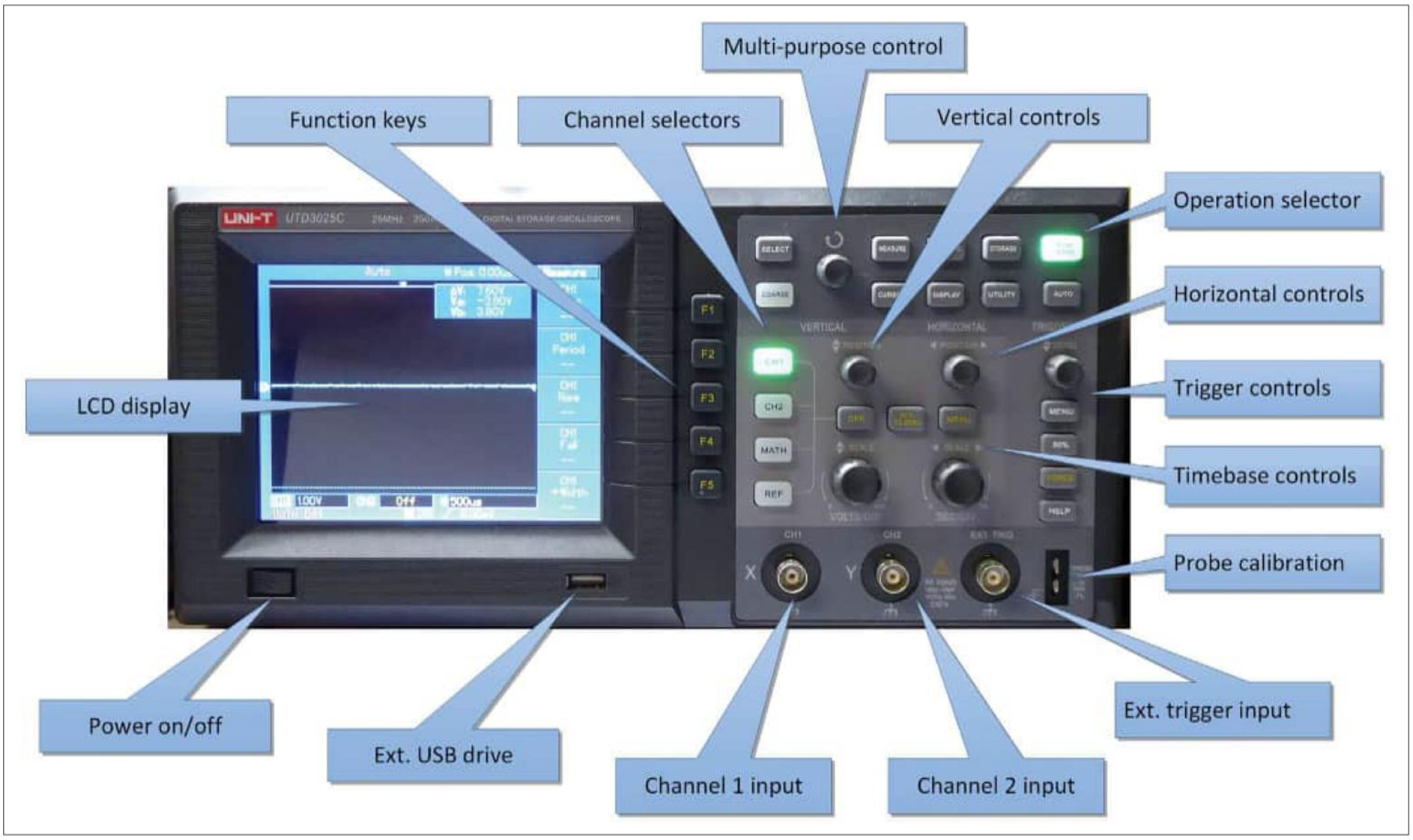


Fig.12.2. Controls and adjustments available on a typical low-cost dual-channel DSO. This 25MHz instrument offers sampling rates of 250MS/s with a resolution of 8-bits and is fitted with a 145mm (5.7-inch) 320×240 LCD.

instruments, such as internal memory and the ability to display information as text.

3. Would you consider a USB instrument as an alternative to a stand-alone 'scope?
USB instruments can be extremely cost-effective but require the services of a host PC or laptop (some can also be used with phones and tablets). If you choose a USB instrument it is important to check that it is compatible with both the hardware and operating system of your host computer. Some USB instruments may only support a limited number of operating systems, so it is worth checking that you have the correct version installed.

4. What display type/size is best for you?
Larger screen sizes are much easier to read (and consequently obtain information) but usually mean a larger instrument. Smaller displays can be cramped and difficult to read but note that if you decide upon a USB 'scope (rather than a free-standing instrument) your display size will be determined by the PC, laptop or phone that you connect it to. Note also that a typical laptop display can be very much larger than the integrated display fitted to a stand-alone bench instrument. The facility to 'screen grab' a display and store it as an image (often in JPG, PNG, or BMP format) can provide you with a useful record for future reference. Many of the images used in this article were produced this way.

5. Do you need an instrument that will accurately display transients (ie, rapid changes in signal levels)? rather than repetitive waveforms?
If yes, you may require sophisticated trigger facilities to detect the change then capture and display it. Repetitive waveforms do not generally require sophisticated triggering. Note that the memory facility found in a DSO will enable you to capture transient events and analyse them later. This can be one of the key reasons for choosing a DSO over a conventional analogue CRT-based instrument.

6. Over what frequency range will you be measuring?
If you only intend to work with audio and low-frequency signals (eg, signals up to 100kHz, or so) you will only require an instrument with a bandwidth of up to about 10MHz. If you need to work with high frequency signals (particularly non-sinusoidal) it may be essential to have an instrument with a bandwidth of 50MHz, or more. As a rule of thumb, the author recommends choosing an instrument with a quoted analogue bandwidth that is at least five-times higher than the highest signal frequency to be measured. Note that cost can be a significant factor here. For example, 10MHz, 25MHz, and 50MHz versions of popular USB 'scopes from Pico technology are currently priced at £115, £185, and £299 respectively. So, if you're not planning to work with high-frequency signals you might opt for the lower bandwidth and save yourself some cash.

7. How many channels do you need?
Most instruments provide two-channel operation, but 'scopes with four or more inputs will allow you to display, measure and compare several signals simultaneously. A useful feature of most dual channel instruments is that they will allow you to display the result of adding (A+B) or subtracting (A-B) two input signals. This latter feature can be invaluable if you need to compare two signals, displaying what is present in one signal that's absent from the other.

8. How much memory do you need?
This is a pertinent question if you are purchasing a digital storage oscilloscope (DSO) rather than a traditional analogue instrument. When making measurements with a DSO your 'scope will effectively be capturing only a small portion of the signal that you've applied to it. How much of the signal gets captured depends on how much memory is available to store the data. DSO memory buffers range considerably in capacity, from about 8K data samples in budget instruments to around 128M data samples in more expensive DSOs.

9. What resolution and sampling rate do you need?
This is another important question if you are intending to purchase a DSO.

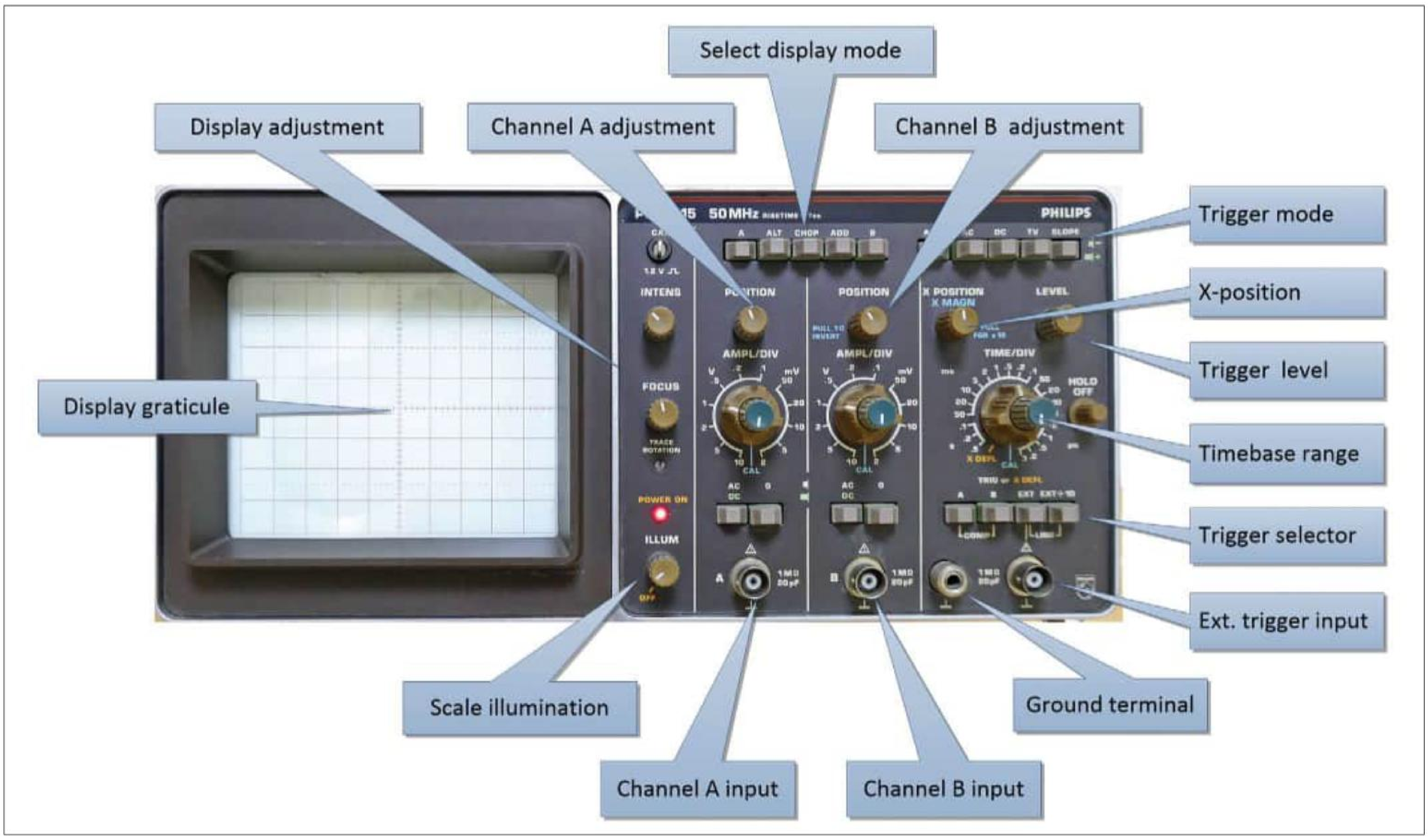


Fig.12.3. Controls and adjustments available on a typical dual-channel CRT-based analogue oscilloscope. This elderly general-purpose 'scope is typical of those available at bargain prices on the second-hand market. It has a 130mm (5.1-inch). screen.

The resolution of a DSO is expressed in terms of the number of bits used in the conversion of the analogue signal to its digital equivalent. For most purposes a resolution of 12 bits will be adequate – but do remember that sampling rate is also important. To be able to perform a detailed display of high-frequency signals it is important to match bandwidth with the real-time sampling rate. For example, to analyse a signal in an analogue bandwidth of 100MHz, a sampling rate of 500MS/s will produce a display with five points plotted for each cycle of the waveform. A sample rate of 1GS/s would produce double the number of points and the waveform will consequently be drawn more precisely. As a rule of thumb, the author recommends choosing an instrument that supports sampling rates that are at least ten-times greater than the highest expected signal frequency.

10. What range of input voltages do you expect?

Some low-cost instruments may have a limited input voltage range (eg, 50V) and others may only be DC-coupled (the popular Hantek 6022 USB 'scope suffers from both limitations). For automotive and power applications an appropriately rated input is essential with switchable AC and DC coupling. A ×10 probe can also assist by reducing the voltage present at the 'scope's input terminal.

11. Do you need additional features such as the ability to read protocol strings?

Modern DSOs offer a vast array of additional features that can sometimes become invaluable. Some DSOs incorporate serial bus triggering and decoding with supported protocols that can include I^{2}C, SPI, UART, RS-232, CAN, LIN, and the list goes on. Others may have inbuilt digital waveform generators capable of producing arbitrary test signals. While none of these features are likely to be deal breakers to new users, they may at some point be something that you wouldn't want to be without.

How easy is it to use?

If you've not used a 'scope before the controls and adjustments can be baffling. In Fig.12.2 and Fig.12.3 we've shown the controls and adjustments found on a typical mid-range DSO and CRT-based instruments. You might find that it takes some time to get familiar with these, but the investment in time and effort can be very rewarding as it will allow you to get the very best out of the instrument. The procedure and adjustments differ according to the type of waveform being investigated and whether the instrument is being used in single- or dual-channel mode (the latter displaying two waveforms simultaneously).

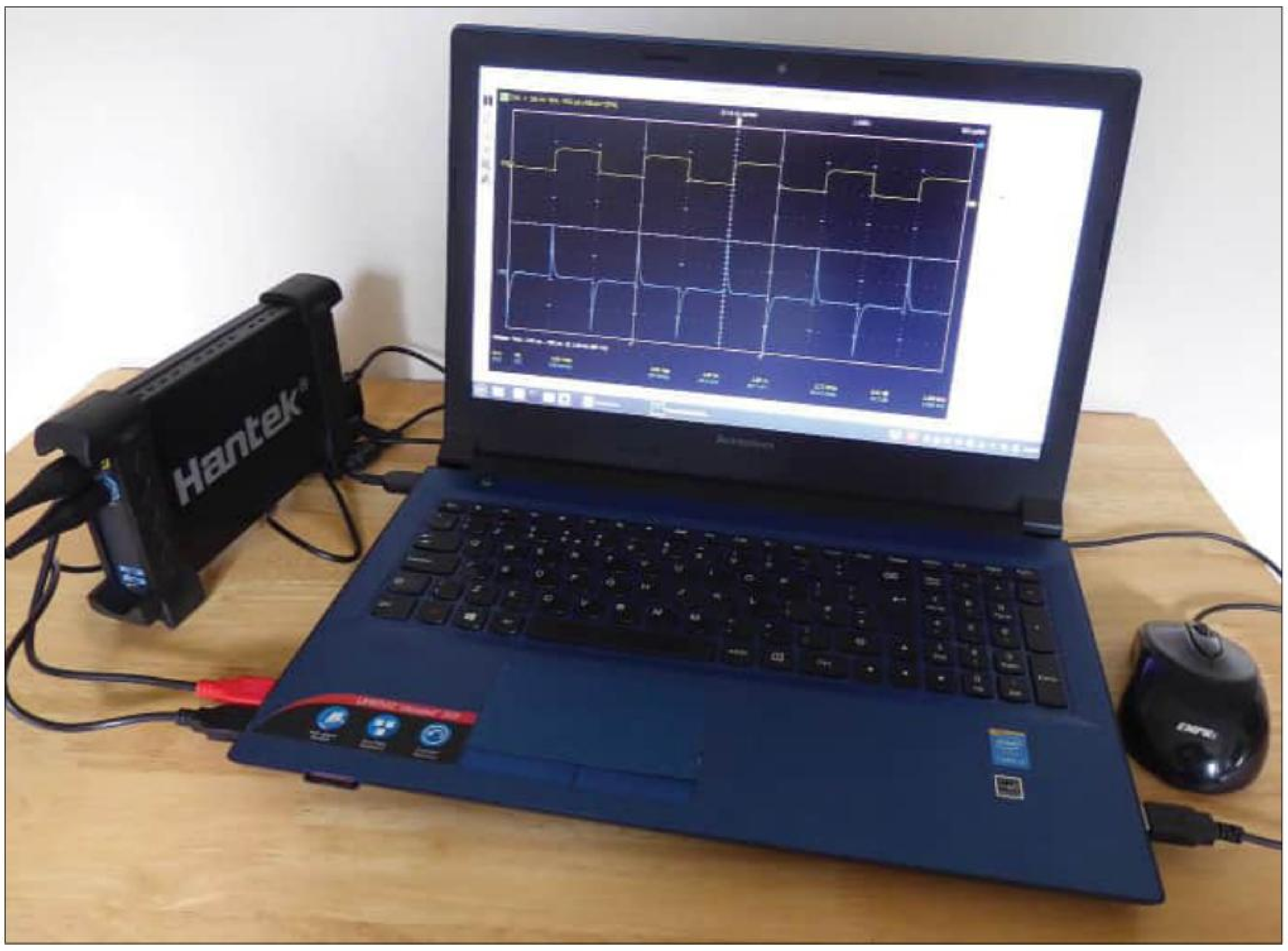


Fig.12.4. The budget-priced Hantek 6022BE USB DSO used in conjunction with the author's Linux-based laptop. Note the size and clarity of the display.

As can be seen from Fig.12.2 and Fig.12.3, many of the controls present in a DSO have direct equivalents in a CRT-based instrument. Others, such as menu selection and function keys are unique to the DSO. The usual input and trigger selectors are present, so too is the trigger level control. As with a CRT-based instrument, the timebase control is marked in terms of 'time per division', – however, it may also be possible to configure a DSO in terms of 'time per scan' which may make more sense for some measurements. For a dual-channel DSO you can select which of the two channels (or both) to display on the screen – but, unlike a CRT-based analogue 'scope, each of the two channel traces can be displayed in a different colour. Voltage ranges are selected in much the same way as for a CRT-based 'scope, but an auto-ranging facility may also be included. This option can be particularly useful if you need to switch between different input signals.

Getting started

The following sequence of adjustments will provide you with a starting point for making most oscilloscope measurements. Many DSOs incorporate automated measurement tools that can simplify most common tasks (such as voltage, time or frequency measurement) but knowing how to make manual measurements can help you understand and check the validity of automatic measurements.

1. Connect power and switch the instrument 'on'. Most DSOs will take a short time to complete their initial boot sequence (after which a display will appear). Older CRT-based instruments may take longer for the CRT's heater to warm and produce a display. On a CRT-based instrument you may need to set the brightness and focus controls to mid-position and then adjust them for a reasonably bright and properly focussed trace (this should appear as a horizontal line on the screen).
2. Adjust the vertical and horizontal shift controls so that the trace aligns with the graticule/scale. Note that older CRT-based 'scopes often have a trace rotation adjustment that can be useful if the CRT is out of alignment with an external scale.
3. Set the horizontal (timebase) range to 1ms/div and the variable timebase control to the calibrate (CAL) position.
4. Set the vertical (voltage) range to 1V/div and the variable voltage control to the calibrate (CAL) position.
5. Set the input coupling to AC.
6. Set the trigger source to Channel A and the trigger mode to AUTO.
7. Attach probes to the instrument and, if necessary, check and compensate the probes (see later).
8. Connect the probe ground clip to common/ground on the circuit under investigation and then connect the probe tip to the desired test point.
9. Re-adjust the time and voltage range controls to obtain a display for measurement purposes. The trigger controls can also be adjusted at this stage to select the exact point on the waveform at which triggering occurs,

Worth considering?

Having decided on the type of oscilloscope that most closely meets your needs, here is a shortlist of popular instruments of each type that are worth considering.

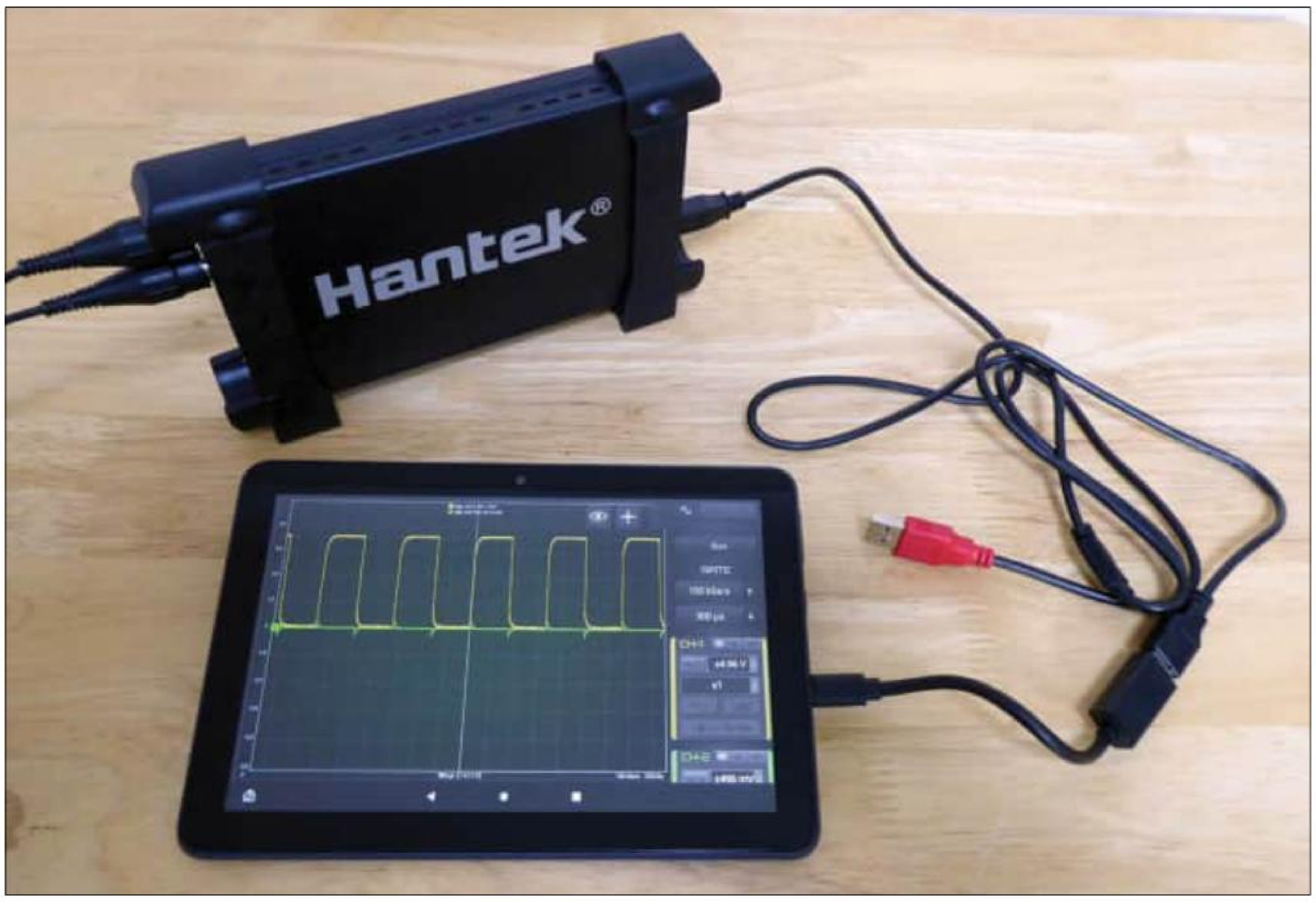


Fig.12.5. The budget-priced Hantek 6022BE USB DSO used in conjunction with an Android-based Amazon Fire tablet. This set-up is ideal for portable use.

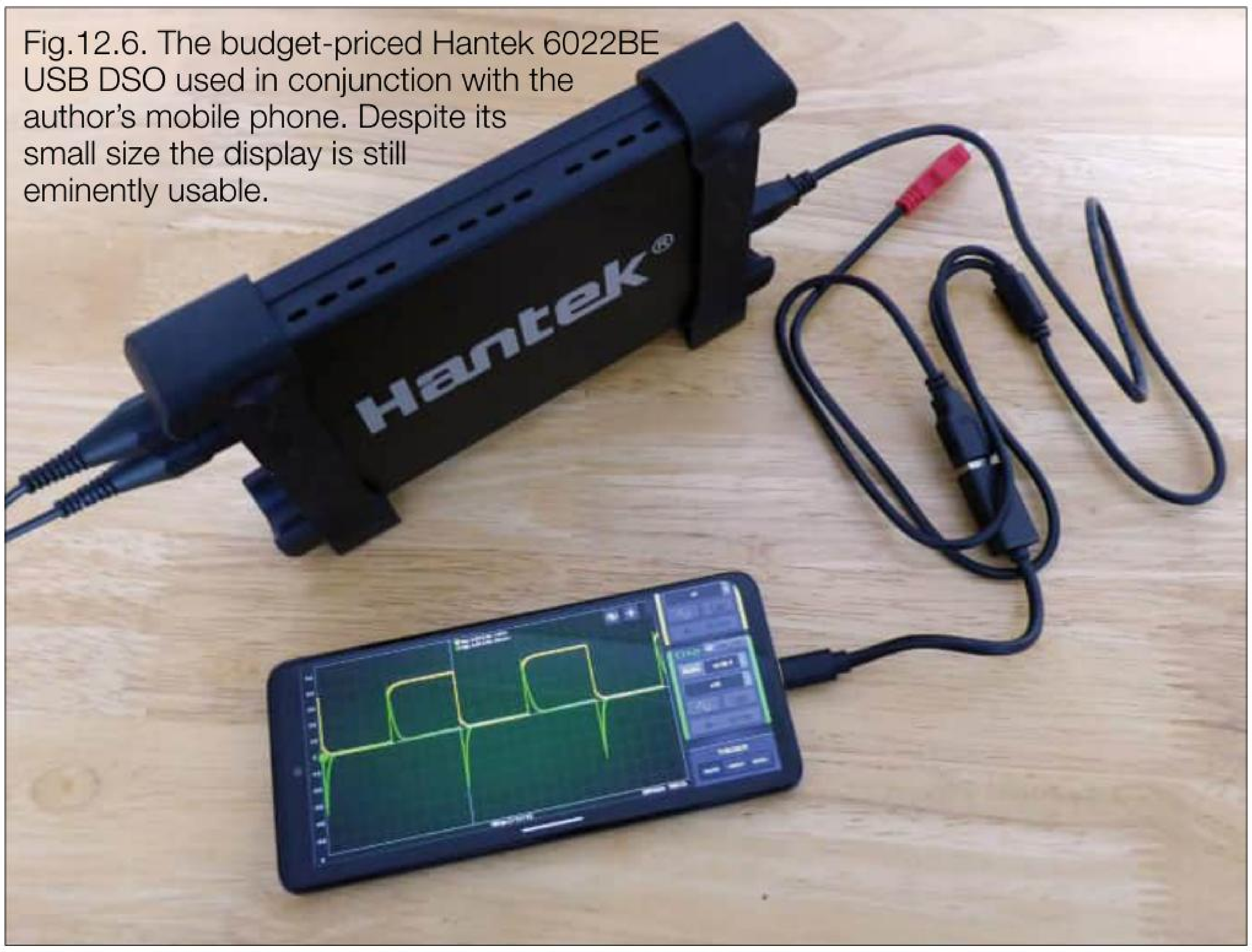


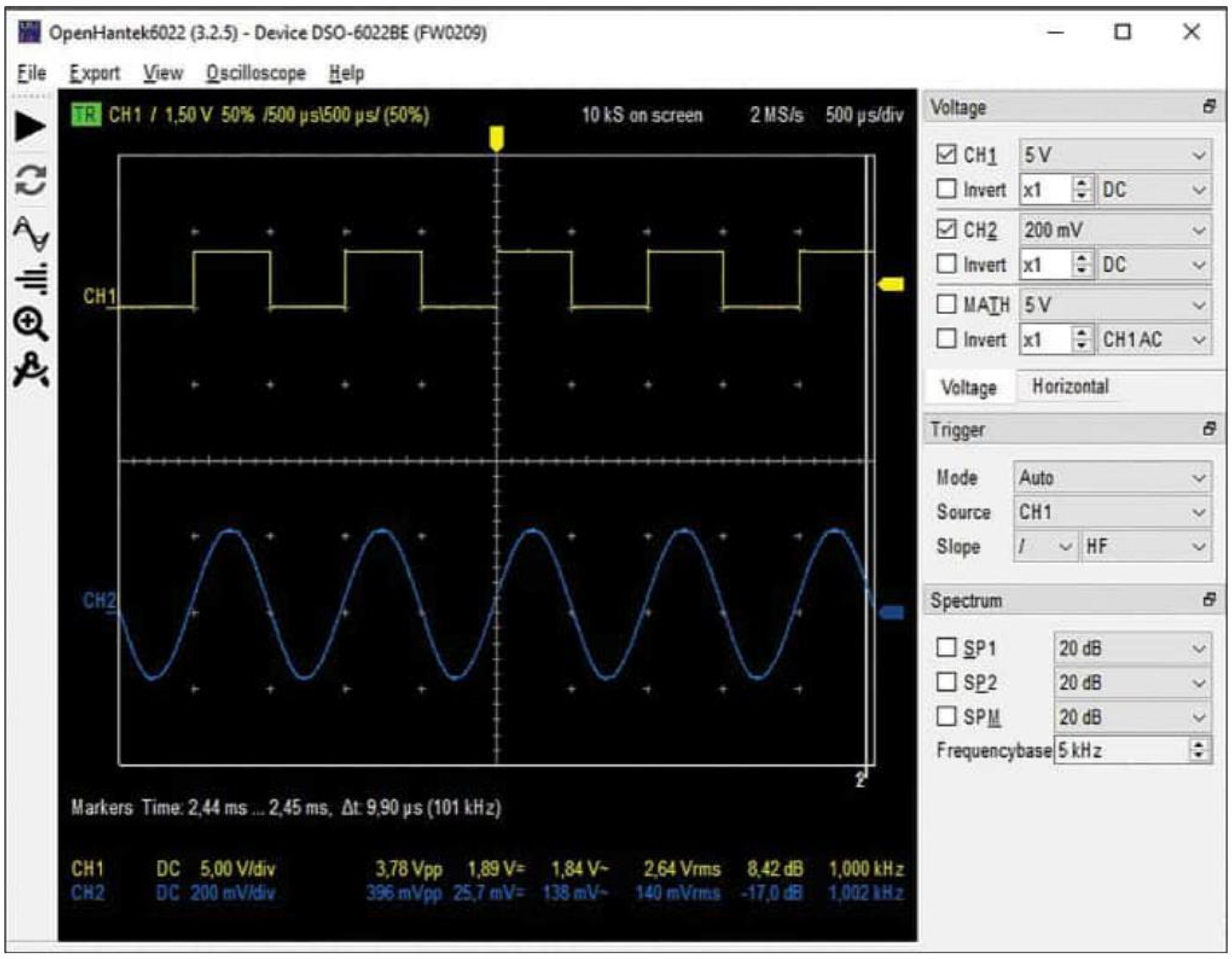
Fig.12.6. The budget-priced Hantek 6022BE USB DSO used in conjunction with the author's mobile phone. Despite its small size the display is still eminently usable.

For those on a strict budget or those who might prefer a basic CRT display, there are numerous instruments to choose from. We've included several of those in the list that follows.

Rigol DS1000Z series (stand-alone DSO)
This is a series of 2- and 4-channel oscilloscopes offering real-time sampling rates of up to 1GS/s and analogue bandwidths from 50MHz to 200MHz. S-versions also have a 2-channel waveform generator and Plus-versions add 16 digital input channels. Prices range from about £360 to around £480.

Hantek DSO2000 series (stand-alone DSO)
Hantek's DSO2000 series oscilloscopes offer an excellent specification for a very modest outlay. They offer two channels with bandwidths of either 100MHz or 50MHz and a maximum sample rate of 1GS/s. Vertical sensitivity ranges from 2mV/div to 10V/div. Two of the instruments in this range (DSO2D10 and DSO2D15) also have a 25MHz in-built arbitrary waveform generator (AWG). The instruments also provide digital readout of voltage and frequency with serial bus data decode and protocol analysis. Prices range from £175 to £275.

Hantek 6022BE (USB DSO)
This low-cost 20MHz USB instrument is designed for the entry-level user. It offers dual-channel operation and is supplied with probes and accessories. Perhaps, not surprisingly for this popular budget 'scope, there is a large user-base, and several free and open-source packages are available as an alternative to the software supplied by Hantek. The 'scope can be used with different types of computer and different operating systems, including Windows, Linux, and Android (see Figs.12.4 to 12.9). The 6022BE has several important limitations, the most noteworthy is that the input is DC coupled and the input voltage range is limited to a maximum of 30V without external attenuation. An AC input modification has been published, see *Going further* for details. The 6022BE is available from several on-line sources at prices ranging from around £60 to £100.



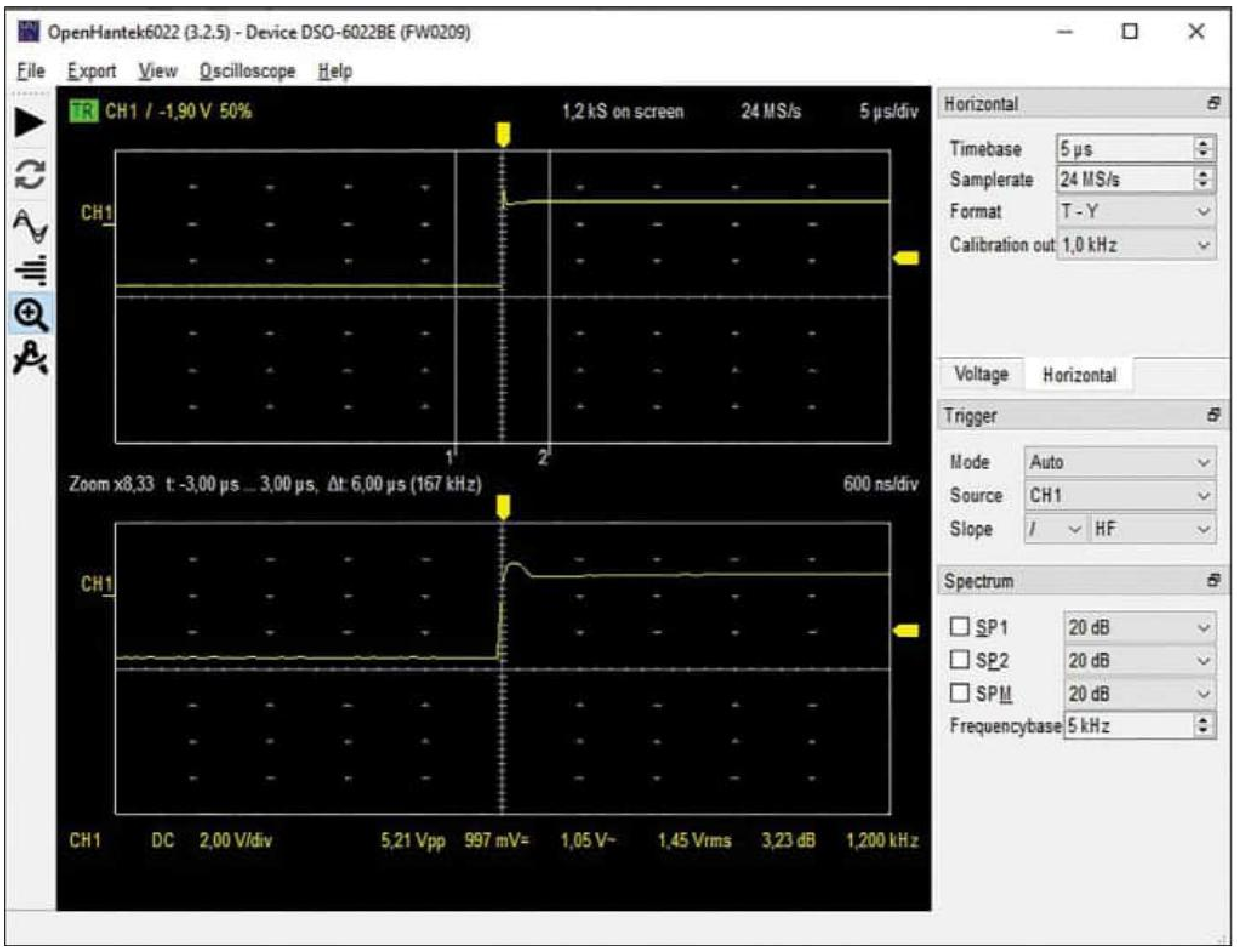
```
TR CH1 / 1,50 V 50% /500 µs\500 µs/ (50%)     10 kS on screen     2 MS/s     500 µs/div

Markers Time: 2,44 ms ... 2,45 ms,  Δt: 9,90 µs (101 kHz)

CH1   DC   5,00 V/div     3,78 Vpp    1,89 V=   1,84 V~   2,64 Vrms    8,42 dB   1,000 kHz
CH2   DC  200 mV/div    396 mVpp   25,7 mV=   138 mV~   140 mVrms  -17,0 dB   1,002 kHz
```

Voltage
☑ CH1 5 V
☐ Invert x1 DC
☑ CH2 200 mV
☐ Invert x1 DC
☐ MATH 5 V
☐ Invert x1 CH1 AC

Trigger
Mode Auto
Source CH1
Slope / HF

Spectrum
☐ SP1 20 dB
☐ SP2 20 dB
☐ SPM 20 dB
Frequencybase 5 kHz

Fig.12.7. The freely available OpenHantek software is an excellent alternative to Hantek's own software. Note the useful information that appears as text at the bottom of the screen.



OpenHantek6022 (3.2.5) - Device DSO-6022BE (FW0209)

TR CH1 / -1,90 V 50%   1,2 kS on screen   24 MS/s   5 µs/div

Zoom x8,33 t: -3,00 µs ... 3,00 µs, Δt: 6,00 µs (167 kHz)   600 ns/div

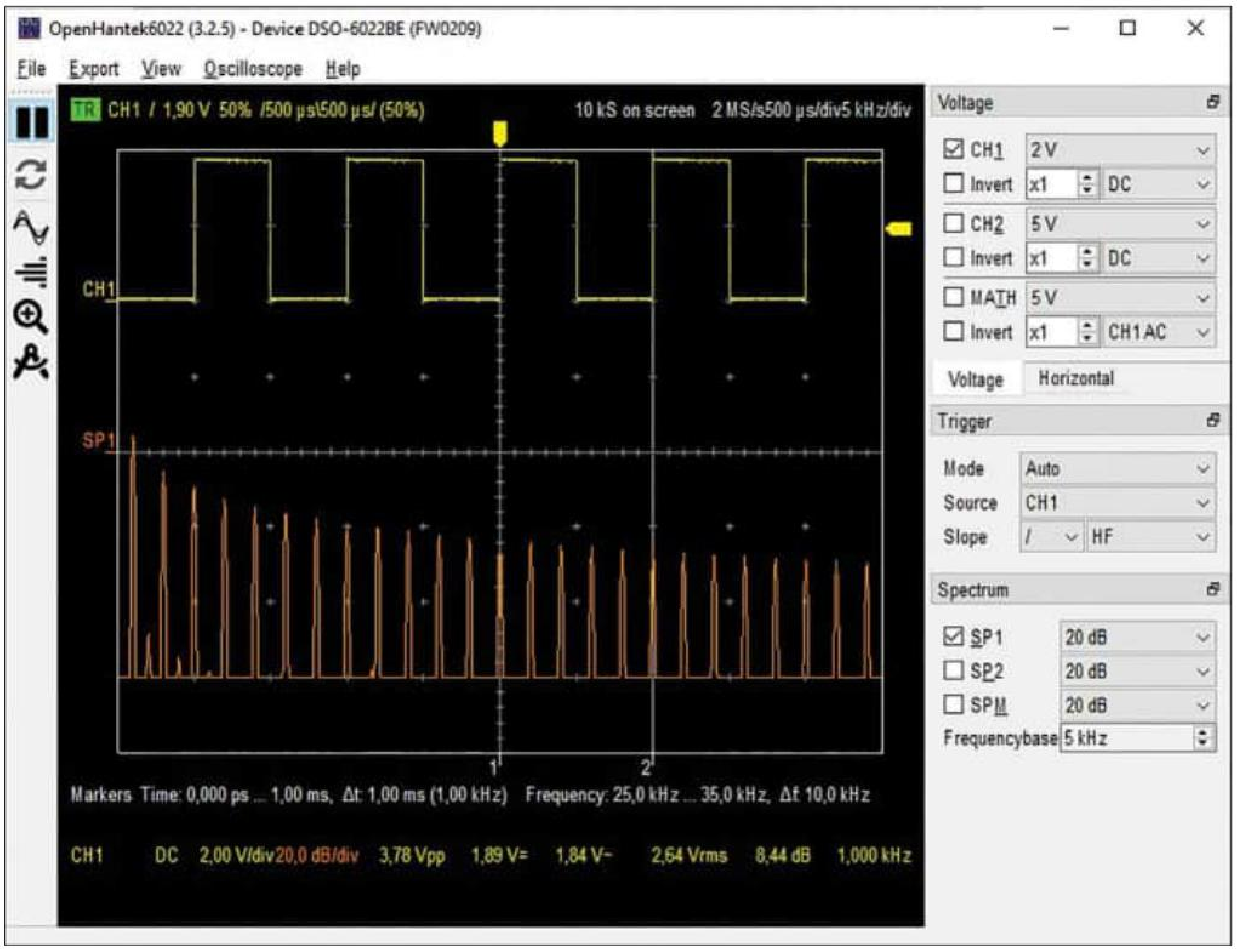
CH1   DC   2,00 V/div   5,21 Vpp   997 mV=   1,05 V~   1,45 Vrms   3,23 dB   1,200 kHz

Horizontal
Timebase: 5 µs
Samplerate: 24 MS/s
Format: T - Y
Calibration out: 1,0 kHz

Trigger
Mode: Auto
Source: CH1
Slope: / HF

Spectrum
☐ SP1: 20 dB
☐ SP2: 20 dB
☐ SPM: 20 dB
Frequencybase: 5 kHz

Fig.12.8. The OpenHantek software incorporates a useful zoom facility that will allow you to inspect a small portion of the applied signal (in this case a 600ns window at the leading edge of a pulse).



OpenHantek6022 (3.2.5) - Device DSO-6022BE (FW0209)

TR CH1 / 1,90 V 50% /500 µs\500 µs/ (50%)   10 kS on screen   2 MS/s500 µs/div5 kHz/div

Markers Time: 0,000 ps ... 1,00 ms, Δt: 1,00 ms (1,00 kHz)   Frequency: 25,0 kHz ... 35,0 kHz, Δf: 10,0 kHz

CH1   DC   2,00 V/div20,0 dB/div   3,78 Vpp   1,89 V=   1,84 V~   2,64 Vrms   8,44 dB   1,000 kHz

Voltage
☑ CH1: 2 V
☐ Invert: x1 DC
☐ CH2: 5 V
☐ Invert: x1 DC
☐ MATH: 5 V
☐ Invert: x1 CH1 AC

Trigger
Mode: Auto
Source: CH1
Slope: / HF

Spectrum
☑ SP1: 20 dB
☐ SP2: 20 dB
☐ SPM: 20 dB
Frequencybase: 5 kHz

Fig.12.9. The OpenHantek software offers spectral analysis of the signal under investigation. In this case you can see the odd harmonic components present in a 1kHz square wave signal (note that the second and fourth harmonic can also be seen at low level).

PicoScope 2200A series (USB DSO)
This is 2-channel 200MHz USB 'scope operates with sampling rates of up to 1GS/s. The supplied software (available for free download) is well supported and easy to use with free updating. The software incorporates various useful features including on-screen waveform measurement data, mask limit testing, serial bus decoding and versatile digital triggering. As an additional bonus, the instrument also incorporates an arbitrary waveform generator (AWG).

Hameg HM203 (second-user analogue 'scope)
Ease of operation and a well-designed front panel layout made this robust low-cost instrument ideal for use in education and training, and this helped ensure its popularity in a wide range of educational contexts. The HM 203 provides basic two-channel operation coupled with a bandwidth of 20MHz. Prices range from around £50 to £100.

Philips PM3215 (second-user analogue 'scope)
Designed for general purpose laboratory and TV/video servicing, this competent 50MHz oscilloscope offers a maximum sensitivity of 2mV/div and a time-base range extending from 0.5s/div to 0.1µs/div. Prices range from about £90 to £180.

Telequipment D83 (second-user analogue 'scope)
The D83 is a versatile laboratory-grade instrument that uses two plug-ins that are also compatible with Telequipment's popular D75, D63 and DM63 oscilloscopes. The available vertical system plug-ins include the V4 dual-trace amplifiers, the V3 single-channel differential amplifier, the V1 single-channel amplifier, and the V5 single-channel amplifier with delay line (primarily intended for the DM63/D63 mainframes that have no built-in delay lines). The instrument offers a sensitivity extending from 5mV/div to 20 V/div and a bandwidth of DC to 50MHz. Prices range from about £60 to £120.

Tektronix 2445B (second-user analogue 'scope)
Developed in the late 1980s, the 2445B was one of the last of a long series of well specified analogue oscilloscopes developed by Tektronix (a market leader since 1946). The 2445B offers four channels, a bandwidth of 150MHz and a timebase adjustable to 2ns/div. The instrument provides on-screen cursors for accurate voltage and time measurement with CRT readout of parameters including voltage, time and frequency. Prices range from £150 to around £250.

Soundcard-based 'scopes (various)
This, potentially zero-cost solution, involves using a soundcard (either internal or external) together with oscilloscope software (see *Going further* for details). Recommended packages are Christian Zeitnitz's Soundcard Scope (see Fig.12.10) and Virtins Multi-Instrument (see Fig.12.11). Please note the restrictions on performance that we mentioned earlier (see Fig.12.12 for an example of poor square wave performance obtained from soundcard instruments).

A word about probes

Earlier we mentioned the importance of using a purpose-designed 'scope probe when taking accurate measurements. If your 'scope doesn't come with a set of probes you will need to acquire one or more switchable ×1 and ×10 probes (see Fig.12.13). Note that a ×10 probe provides an *attenuation* of 10 times.

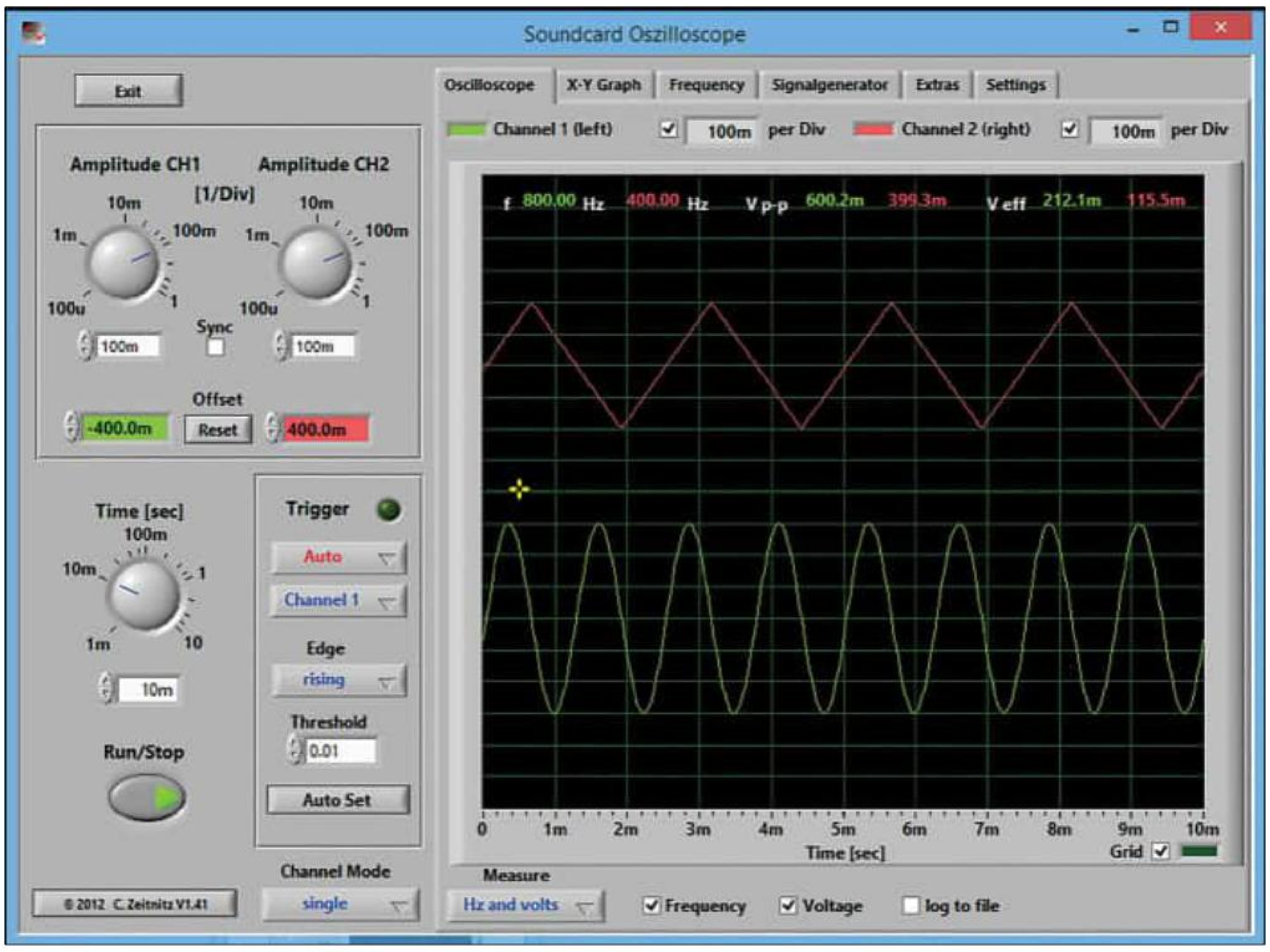
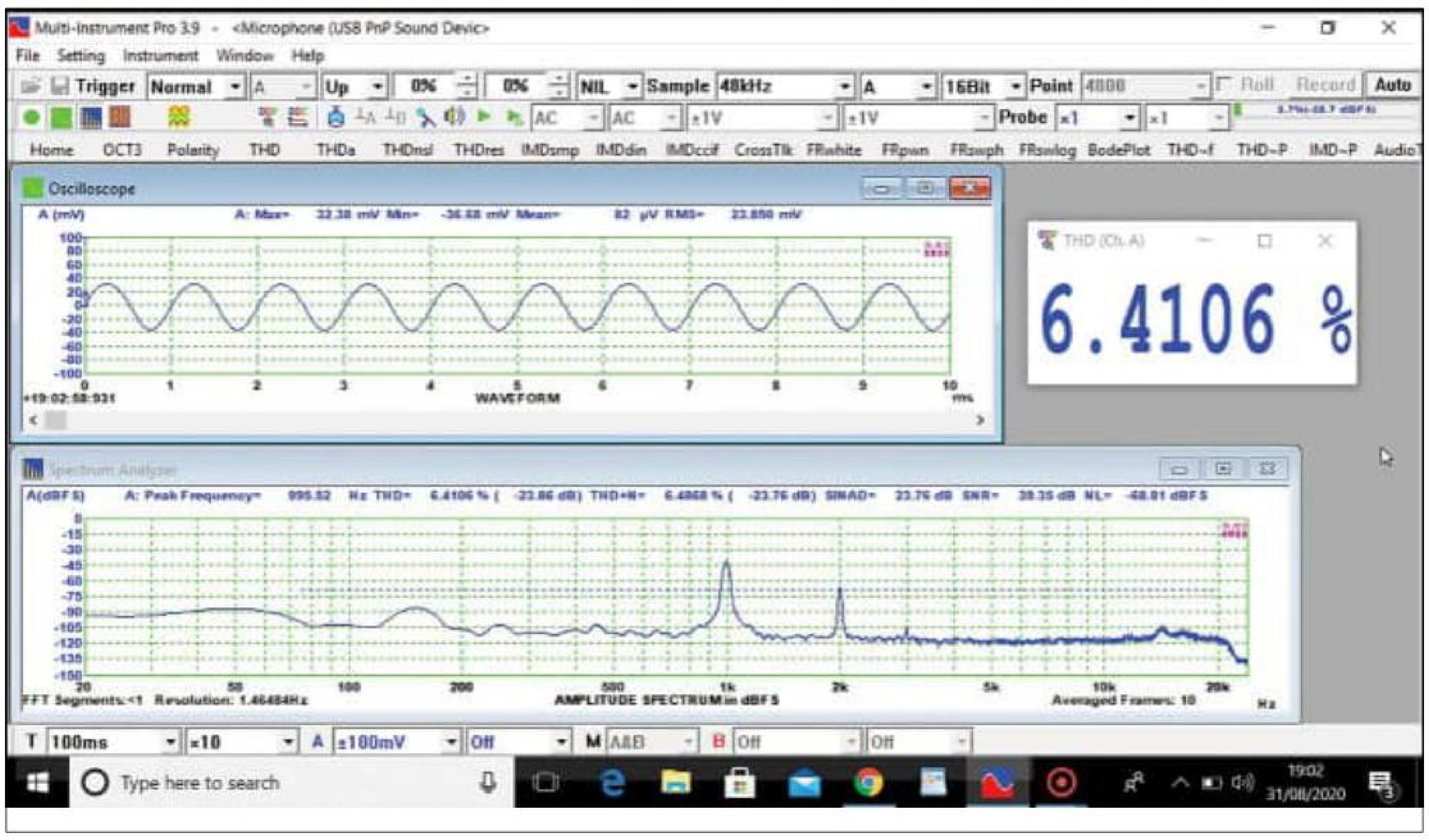

Soundcard Oszilloscope
Exit
Oscilloscope
X-Y Graph
Frequency
Signalgenerator
Extras
Settings
Channel 1 (left)
100m per Div
Channel 2 (right)
100m per Div
Amplitude CH1
Amplitude CH2
[1/Div]
Sync
Offset
-400.0m
Reset
400.0m
f 800.00 Hz 400.00 Hz
V p-p 600.2m 399.3m
V eff 212.1m 115.5m
Time [sec]
Trigger
Auto
Channel 1
Edge
rising
Threshold
0.01
Auto Set
Run/Stop
Channel Mode
single
© 2012 C.Zeitnitz V1.41
Measure
Hz and volts
☑ Frequency
☑ Voltage
☐ log to file
Grid ☑

Fig.12.10. A typical soundcard 'scope display produced by Christian Zeitnitz's excellent Soundcard Scope software. The display shows low-frequency triangle and sine waveforms (red and green traces respectively).


Multi-Instrument Pro 3.9 - <Microphone (USB PnP Sound Device)>
File Setting Instrument Window Help
Trigger Normal
Sample 48kHz
16Bit
Oscilloscope
WAVEFORM
THD (Ch. A)
6.4106 %
Spectrum Analyzer
A: Peak Frequency= 995.52 Hz THD= 6.4106 % (-23.86 dB) THD+N= 6.4868 % (-23.76 dB) SINAD= 23.76 dB SNR= 38.35 dB NL= -68.81 dBFS
AMPLITUDE SPECTRUM in dBFS
Type here to search
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31/08/2020

Fig.12.11. Virtin's powerful and comprehensive Multi-Instrument software being used with a PC soundcard to investigate and measure the total harmonic distortion (THD) present in a 1kHz sinewave signal.

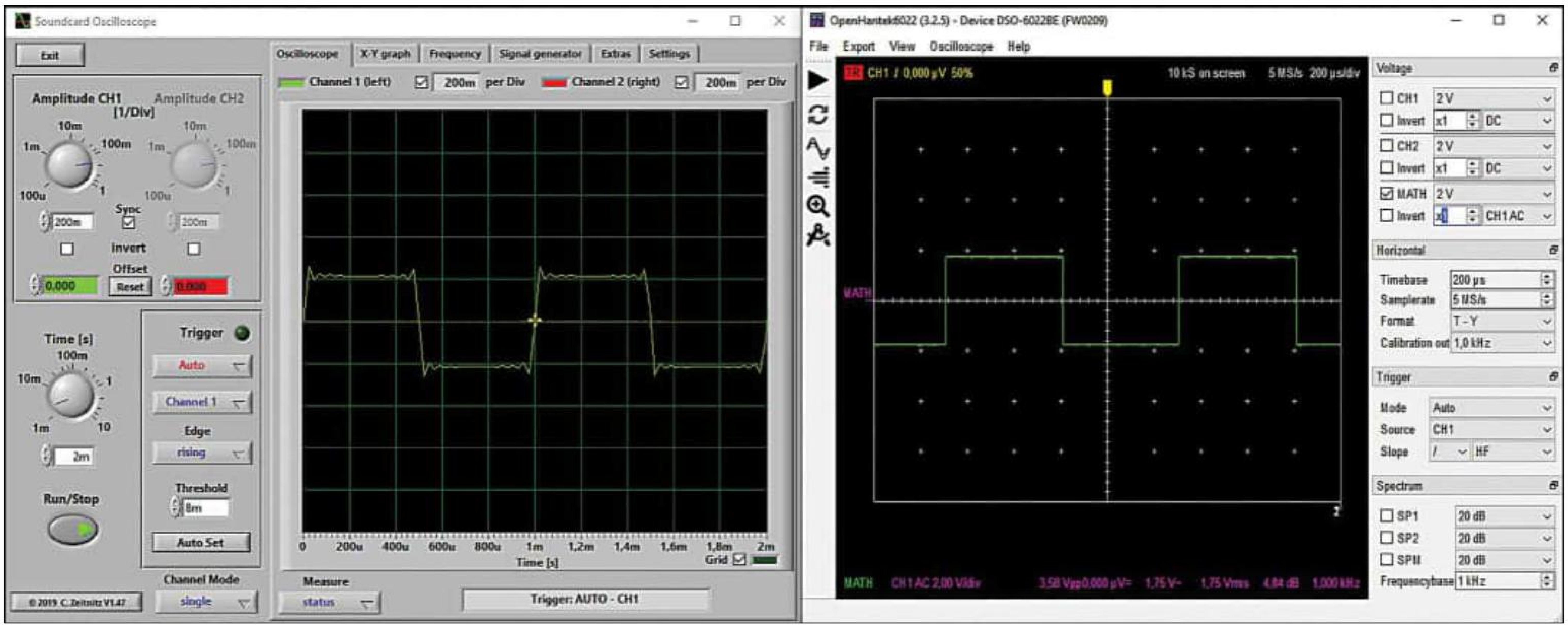
It is important to be aware that 'scope probes are designed to be matched to a particular instrument and require initial calibration before use. It is also worth checking probe compensation every time you set up your 'scope. This can usually be carried out quickly and easily using the internal square wave reference source that's available on most instruments. If you don't have a calibration source, you can use a square wave source or use the handy 'scope calibrator described in our *Get Testing Teach-In 9* series (see *Going further* for details).

An important requirement of a 'scope is that it should faithfully reproduce signals and pulses of fast duration and that it should not load the circuit to which it is connected. The standard input resistance of most 'scopes is 1MΩ, but appearing in parallel with this is a small (stray) capacitance of around 20pF, as shown in Fig.12.14(a). Note that this shunt capacitance also appears in parallel with that of an input cable and this can be appreciable (a typical 50Ω coaxial cable has a capacitance of around 100pF per metre).

Fig.12.14(b) shows the basic arrangement of an uncompensated ×10 probe. A close-tolerance series resistor of 9MΩ forms an attenuator in conjunction with the 1MΩ input resistance of the 'scope. The probe tip then imposes a load of 10MΩ rather than the 1MΩ of the 'scope alone. The unfortunate consequence of this arrangement is that the 9MΩ probe tip resistance forms a low-pass filter with the capacitance of the cable (C_C) acting in parallel capacitance with the nominal 20pF input capacitance of the 'scope. This severely reduces the high frequency response of the 'scope and probe combination.

Compensation can be achieved in various ways to improve the frequency response. Fig.12.14(c) shows how a

Fig.12.12. (below) Comparison of a soundcard 'scope (left) with a DSO (right) displaying a 1kHz square wave signal. Of particular note is the rise and fall times displayed by the soundcard instrument.


Soundcard Oscilloscope
Exit
Oscilloscope
X-Y graph
Frequency
Signal generator
Extras
Settings
Channel 1 (left)
200m per Div
Channel 2 (right)
200m per Div
Amplitude CH1
Amplitude CH2
[1/Div]
Sync
Invert
Offset
0.000
Reset
Time [s]
Trigger
Auto
Channel 1
Edge
rising
Threshold
Auto Set
Run/Stop
Channel Mode
single
© 2019 C.Zeitnitz V1.47
Measure
status
Trigger: AUTO - CH1
Time [s]
Grid ☑
OpenHantek6022 (3.2.5) - Device DSO-6022BE (FW0209)
File Export View Oscilloscope Help
CH1 / 0,000 µV 50%
10 kS on screen
5 MS/s 200 µs/div
Voltage
CH1 2V
CH2 2V
MATH 2V
Invert
DC
CH1 AC
Horizontal
Timebase 200 µs
Samplerate 5 MS/s
Format T - Y
Calibration out 1,0 kHz
Trigger
Mode Auto
Source CH1
Slope / HF
Spectrum
SP1 20 dB
SP2 20 dB
SPM 20 dB
Frequencybase 1 kHz
MATH CH1 AC 2,00 V/div
1,75 V~ 1,75 Vrms 4,84 dB 1,000 kHz

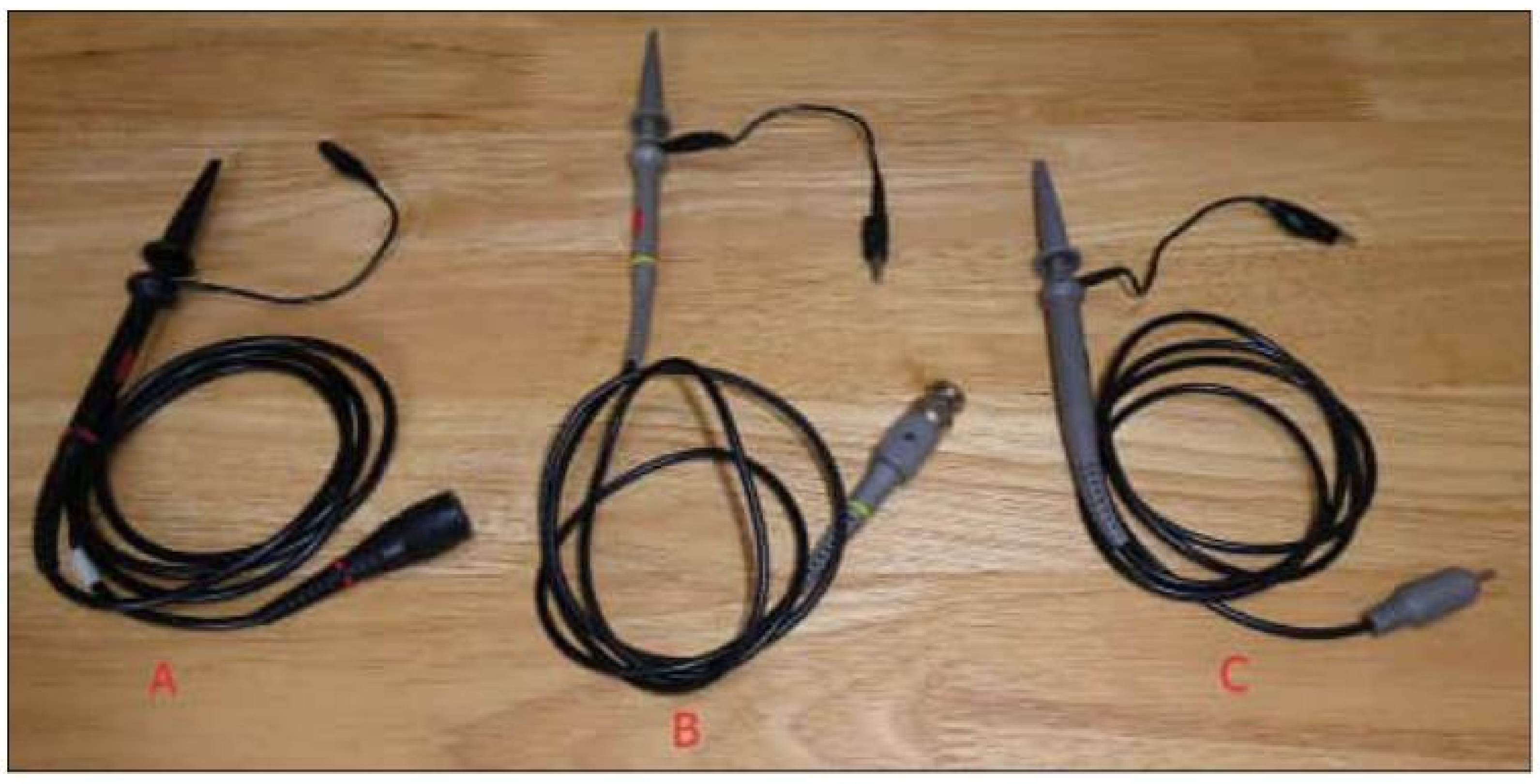


Fig.12.13. Three different oscilloscope probes (see text).

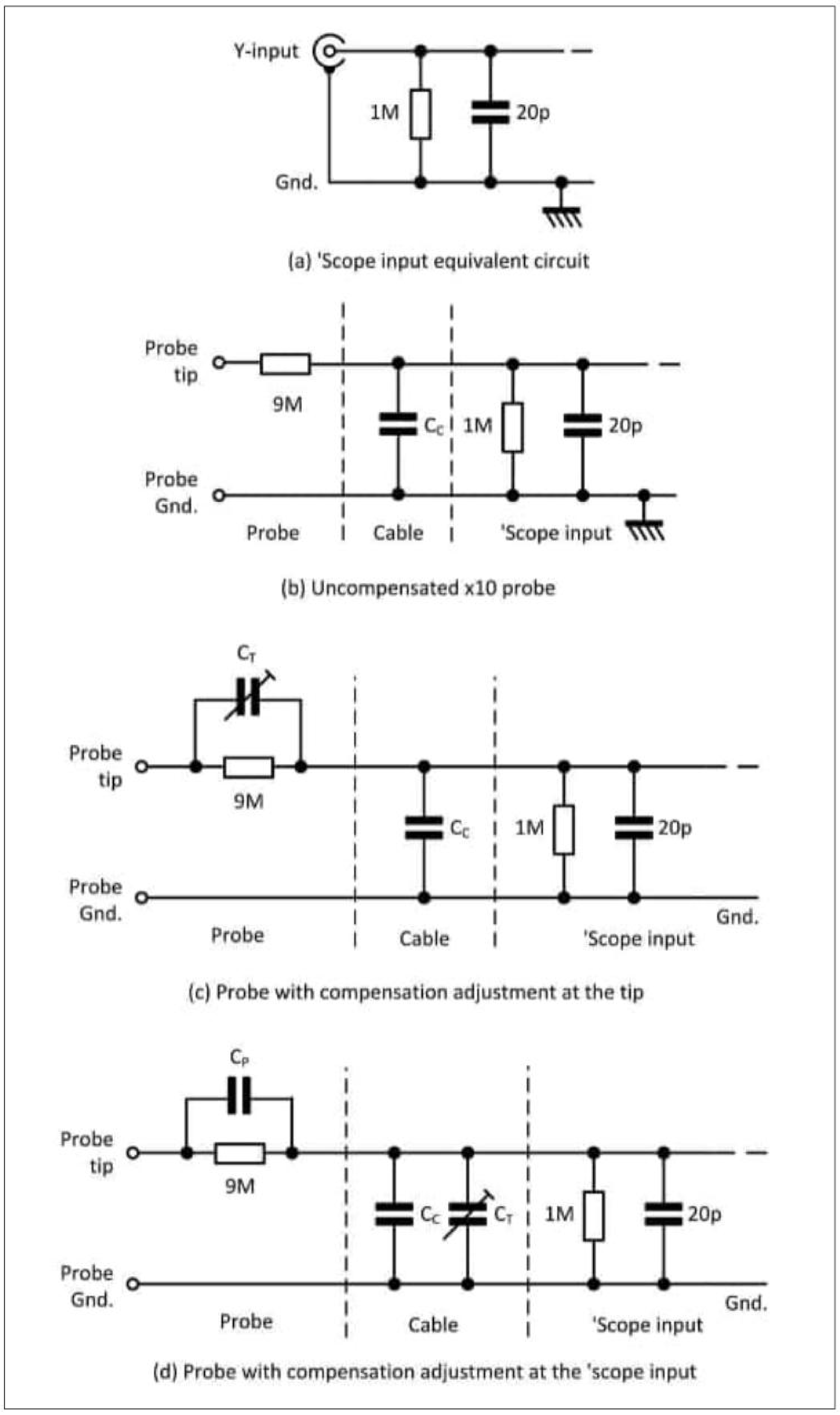

Y-input
1M
20p
Gnd.
(a) 'Scope input equivalent circuit
Probe tip
9M
C_C
1M
20p
Probe Gnd.
Probe
Cable
'Scope input
(b) Uncompensated x10 probe
C_T
Probe tip
9M
C_C
1M
20p
Probe Gnd.
Gnd.
Probe
Cable
'Scope input
(c) Probe with compensation adjustment at the tip
C_P
Probe tip
9M
C_C
C_T
1M
20p
Probe Gnd.
Gnd.
Probe
Cable
'Scope input
(d) Probe with compensation adjustment at the 'scope input

Fig.12.14. Equivalent circuits of the input of an oscilloscope with probes.

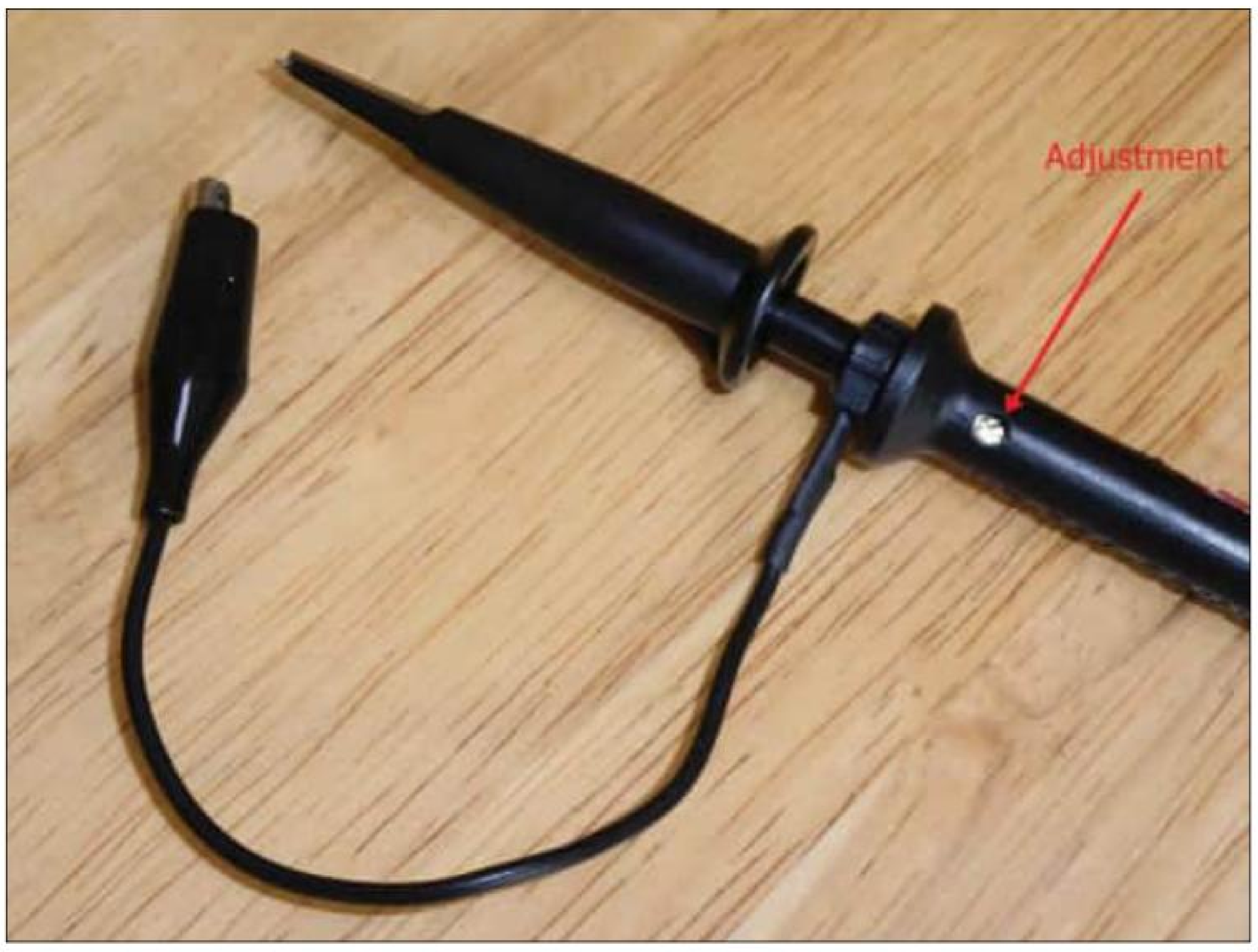

Adjustment

Fig.12.15. Compensation adjustment at the tip of a ‘scope probe.

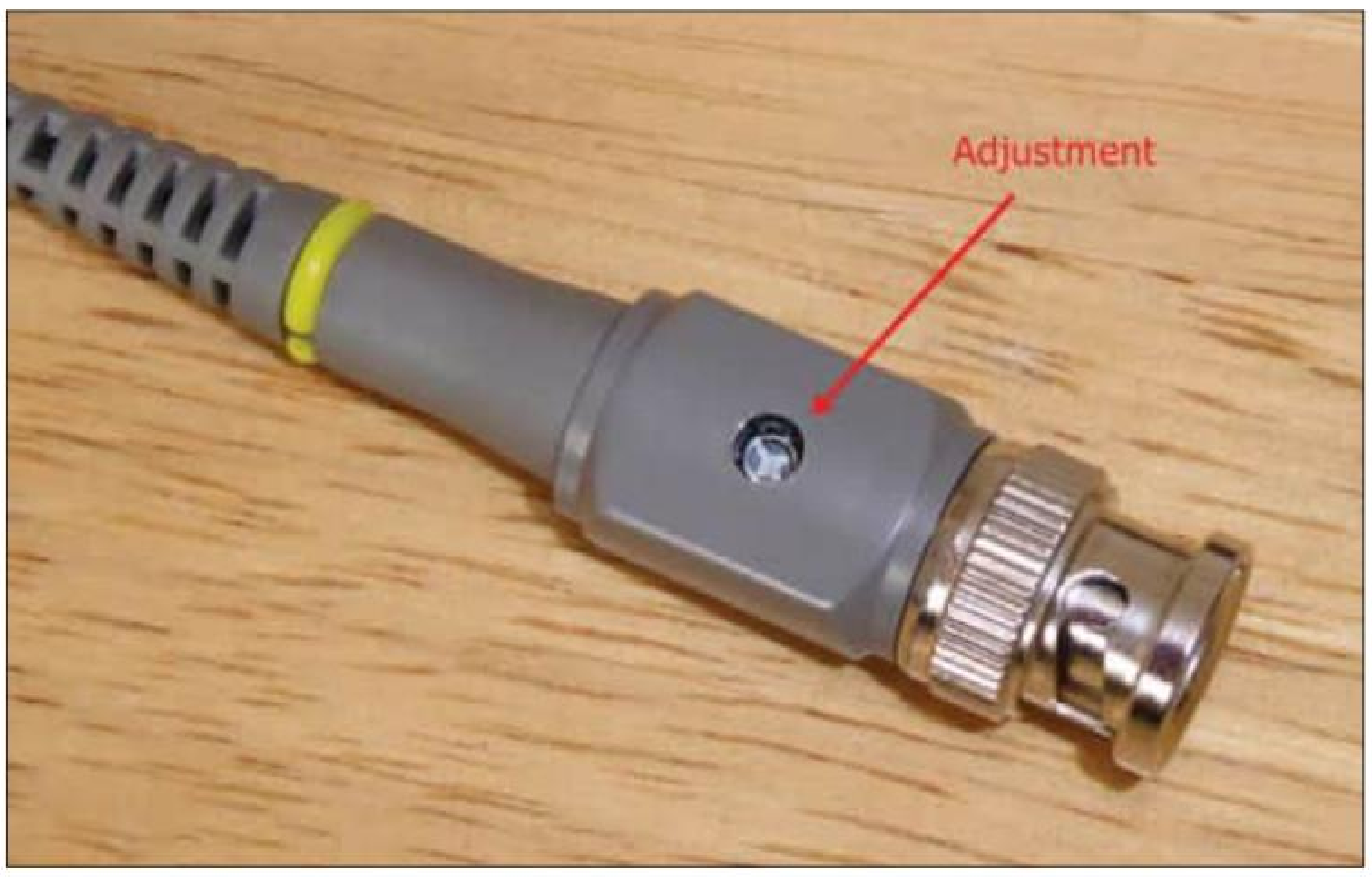

Adjustment

Fig.12.16. Compensation adjustment at the ‘scope input connector.

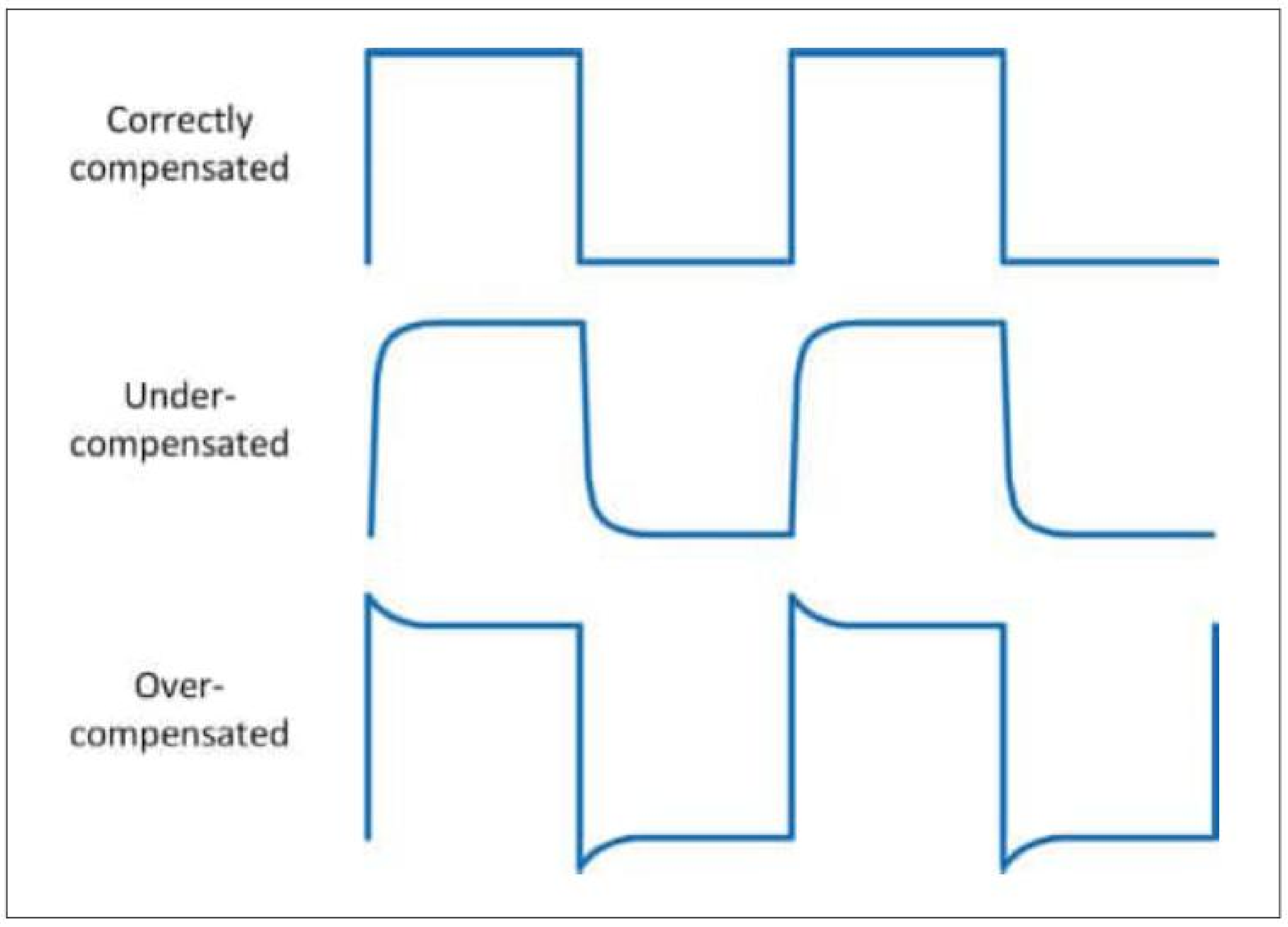


Fig.12.17. Typical waveforms obtained during the compensation of a ‘scope probe.

low-value trimmer capacitor can be introduced in parallel with the 9MΩ probe tip resistor. This arrangement is used in some commercial probes. An alternative arrangement, shown in Fig.12.14(d), uses a fixed capacitor in parallel with the 9MΩ probe tip input resistor and a shunt-connected trimmer capacitor fitted at the ‘scope input.

In Fig.12.13, probe A incorporates a compensation adjustment at the tip while the compensation for probe B is adjusted at the ‘scope input connector. Probe C is designed specifically for use with soundcard ‘scopes and does not incorporate any means of compensation. Fig.12.15 shows how the trimmer adjustment is made accessible at a probe’s tip, while Fig.12.16 shows the equivalent adjustment point at a probe’s BNC connector. In either case, after applying a square wave calibrating signal to the probe tip (see *Going further*) the compensating trimmer is simply adjusted for the best square wave (in other words, a square wave with fast rise and fall times and with no discernible overshoot). Fig.12.17 shows typical waveforms obtained during probe compensation.

Going further

This section details a variety of sources that will help you locate the component parts and further information that will allow you to acquire and get the best out of your oscilloscope. It also provides links to relevant underpinning knowledge and manufacturers’ data sheets.

Table 12.1. Going Further with your first ‘scope

Topic	Source	Notes
Oscilloscope theory and practice	A general introduction to the use of oscilloscopes can be found in *Part 2* of *Electronics Teach-In 9*.	*Electronics Teach-In 9* is available direct from *PE* at: **https://bit.ly/pe-eti9**
Oscilloscope suppliers	The web provides numerous oscilloscope manufacturers and suppliers. Those mentioned in this article include: Picotech: **www.picotech.com** Rigol: **www.rigol-uk.co.uk** Hantek: **www.hantek.com**	Oscilloscopes can also be purchased from most large electronic component and test equipment suppliers.
Oscilloscope calibrator	The *Test Gear Project* featured in *Part 2* of *Electronics Teach-In 9* provides an accurate 1kHz square wave calibration signal with an amplitude of 5V and suitably fast rise and fall times. The ‘scope calibrator is small, inexpensive and easily constructed.	*Electronics Teach-In 9* is available direct from *PE* at: **https://bit.ly/pe-eti9**
Soundcard oscilloscopes	Several useful software packages are currently available for free download. Two that can be highly recommended are Soundcard Oscilloscope by Christian Zeitnitz and Virtin’s Multi-Instrument (the latter is also designed to work in conjunction with Virtin’s own range of USB DSO).	Soundcard Oscilloscope is available from: **www.zeitnitz.eu/Scope_en** Virtin’s Multi-Instrument can be downloaded from: **www.virtins.com/index.shtml**
OpenHantek	Conceived as an alternative to the official Hantek DSO software for Linux users, this package now works with Microsoft Windows and Apple macOS. OpenHantek is also compatible with Voltcraft, Darkwire, Protek and Acetech USB DSOs.	OpenHantek software can be downloaded from: **http://openhantek.org/**
Second-user oscilloscopes	Second-user oscilloscopes appear regularly at on-line suppliers and auction sites.	eBay is a good source, but do check sellers’ feedback and postage/carriage charges.
Hantek 6022BE modification	An AC input modification for the Hantek 6022BE can be found at: **https://bit.ly/pe-jul23-han**	This modification is only recommended for those with electronics experience.

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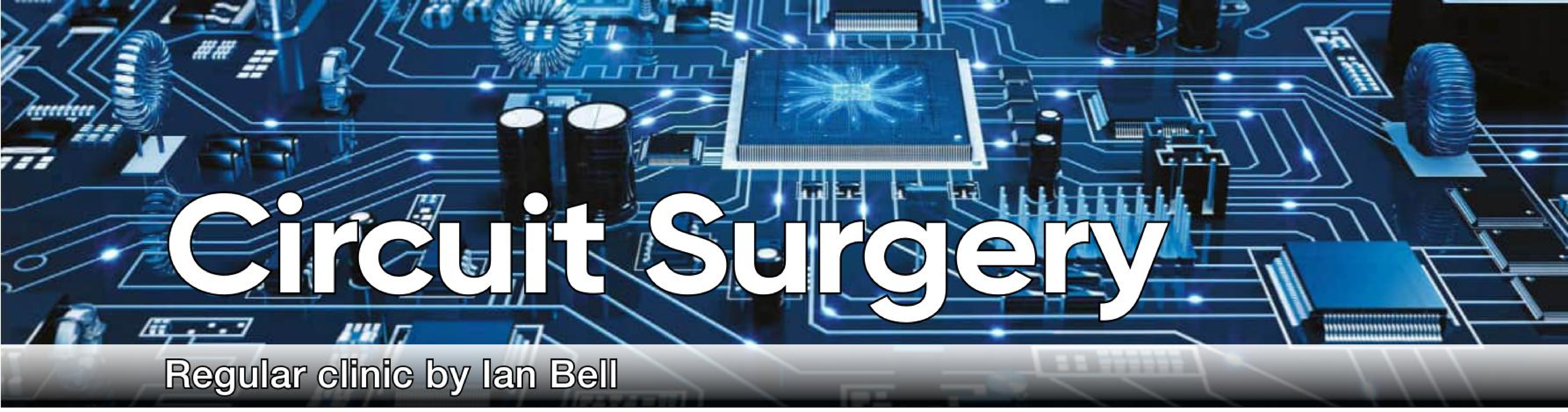


Circuit Surgery

Regular clinic by Ian Bell

Diode power circuits and ideal diode controllers

User Cuocomponents posted the following question on the EEWeb forum. 'Can anyone suggest a device that realises the function of an ideal diode/OR-ing controller with the following parameters: input voltage range from 2.5V, Qualification: AEC-Q100.' Other members replied with a couple of device selections, which was what *Cuocomponents* was looking for. However, inspired by this question, we will look at ideal diode controllers and related circuits in more depth. If you have not heard of these circuits, the headline description is that they are used in power supply circuits for reverse-voltage protection and switching between power sources (eg, batteries and mains DC adapters). Discrete silicon and Schottky diodes can be used in these circuits, but ideal diode controllers provide improved efficiency and performance. The AEC-Q100 referred to in the question is a stress-test qualification for packaged integrated circuits used in automotive applications. Our discussion will introduce ideal diode controllers after covering the more basic circuits they replace; we won't look at details of the AEC-Q100 qualification.

Supply reversal protection

With battery-powered equipment there is always the risk that the user will insert or connect the battery the wrong way round. Unfortunately, circuits may be damaged or destroyed by supply reversal, so steps should be taken to avoid this. It is possible to try to prevent battery reversal by mechanical means. For example, the commonly used PP3 9V battery has different terminal shapes for positive and negative. Despite this, it is still possible to touch the connector to the circuit the wrong way wrong and other batteries (such as AA, C and D types) may be easily inserted the wrong way in certain products. In some systems, for example automotive applications, transient reverse voltages may occur on the supply due to switching inductive loads, which may damage electronic circuits on the same supply. Electronic reverse supply voltage projection is therefore often needed and can be achieved straightforwardly using a diode, as shown in Fig.1.

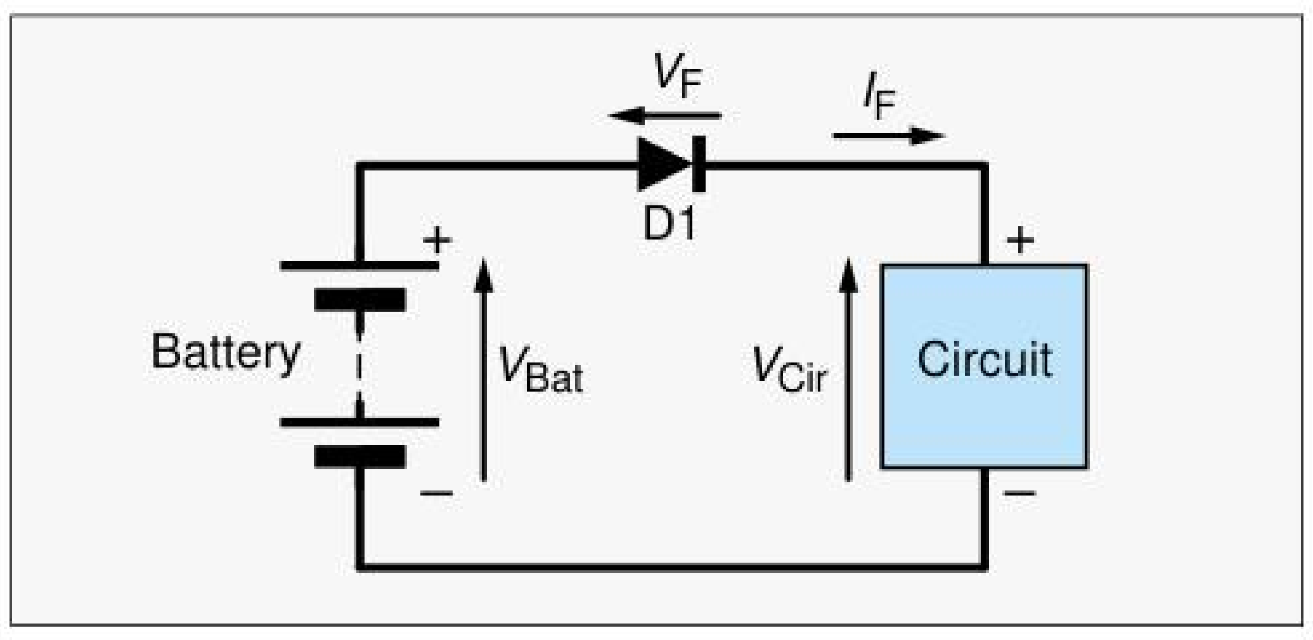


Fig.1. Battery-powered circuit with reverse-voltage protection diode.

Simple diode protection

In the circuit in Fig.1, the battery is shown connected correctly. The diode is forward biased and current flows from the battery to the circuit, which will operate normally. If the battery is reversed the diode becomes reverse biased and will not conduct. There will be a small leakage current, but this will not damage the circuit.

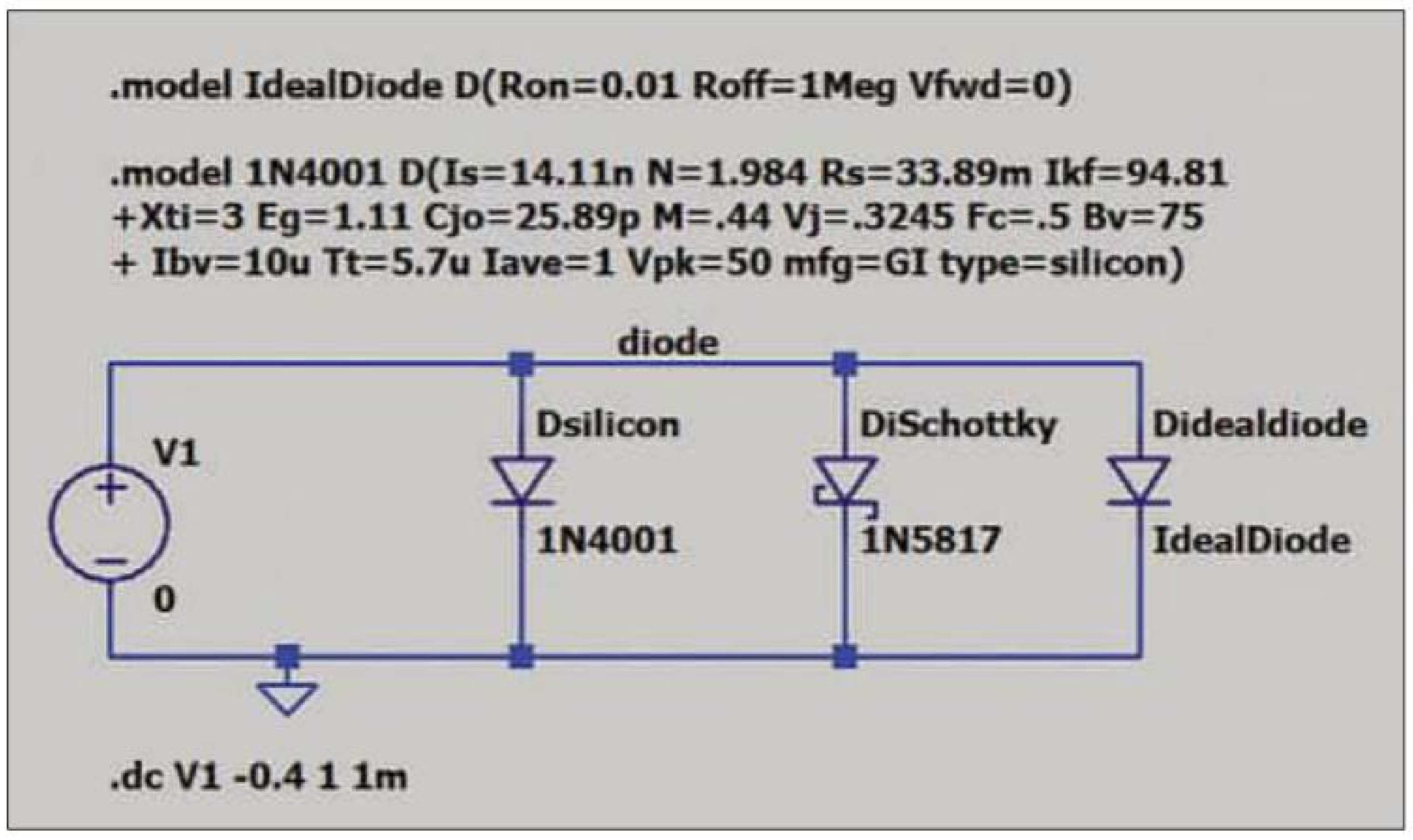
Various things need be to be considered when using the circuit in Fig.1. A diode must be selected with a maximum forward current rating above the supply current of the circuit (I_F is the forward current in Fig.1). The diode's maximum reverse voltage must be larger than the battery voltage, although this is unlikely to be a problem with typical battery-powered circuits and rectifier didoes. For example, the common 1N4001 has a reverse breakdown (blocking) voltage of about 50V and other devices in the 1N400x range have much higher values. The 1N400x range have a maximum forward current of 1A and higher current rectifier diodes are of course available (eg, the 1N540x range at 3A). For circuits with smaller supply currents, signal diodes such as the 1N4148 (300mA maximum forward current), could be used. Signal diodes are not available with such high breakdown voltages as rectifier diodes, but for most battery circuits this will not be an issue.

Adding a reverse protection diode may cause problems. Due to the forward voltage across it (V_F in Fig.1) the diode will dissipate power (equal to $V_F I_F$) which will waste energy from the battery, reducing battery life. The diode also results in a lower voltage across the circuit (V_{Cir}) than the battery voltage (V_{Bat}), specifically $V_{Cir} = V_{Bat} - V_F$. This can also reduce battery life by reducing the amount of battery charge that can be used – more on this later. Both these problems can be reduced if the forward voltage can be reduced. One way to do this is to use a Schottky diode.

Schottky diode protection

Schottky diodes are formed with a metal-semiconductor junction, unlike the junctions of P and N-type silicon used in standard silicon diodes, such as the devices mentioned above. The typical forward voltage for a Schottky diode is 200mV to 300mV, rather than the 600mV to 700mV for silicon PN junction diodes. These typical values are approximate, and relatively low and high currents will produce voltage drops outside these ranges, but under comparable conditions Schottky diodes have a significantly smaller forward voltage.

We can see the difference between Schottky and silicon diodes if we look at their current against voltage (IV) characteristics. From these curves we can find the voltage drop at a given current and hence find the power dissipation in the reverse protection diode when a circuit is operating at a particular supply current. We can plot IV characteristic of diodes using a DC sweep in LTspice – for example, using the setup shown in Fig.2,



```
.model IdealDiode D(Ron=0.01 Roff=1Meg Vfwd=0)

.model 1N4001 D(Is=14.11n N=1.984 Rs=33.89m Ikf=94.81
+Xti=3 Eg=1.11 Cjo=25.89p M=.44 Vj=.3245 Fc=.5 Bv=75
+ Ibv=10u Tt=5.7u Iave=1 Vpk=50 mfg=GI type=silicon)

.dc V1 -0.4 1 1m
```

Fig.2. LTspice circuit to obtain diode IV characteristics.

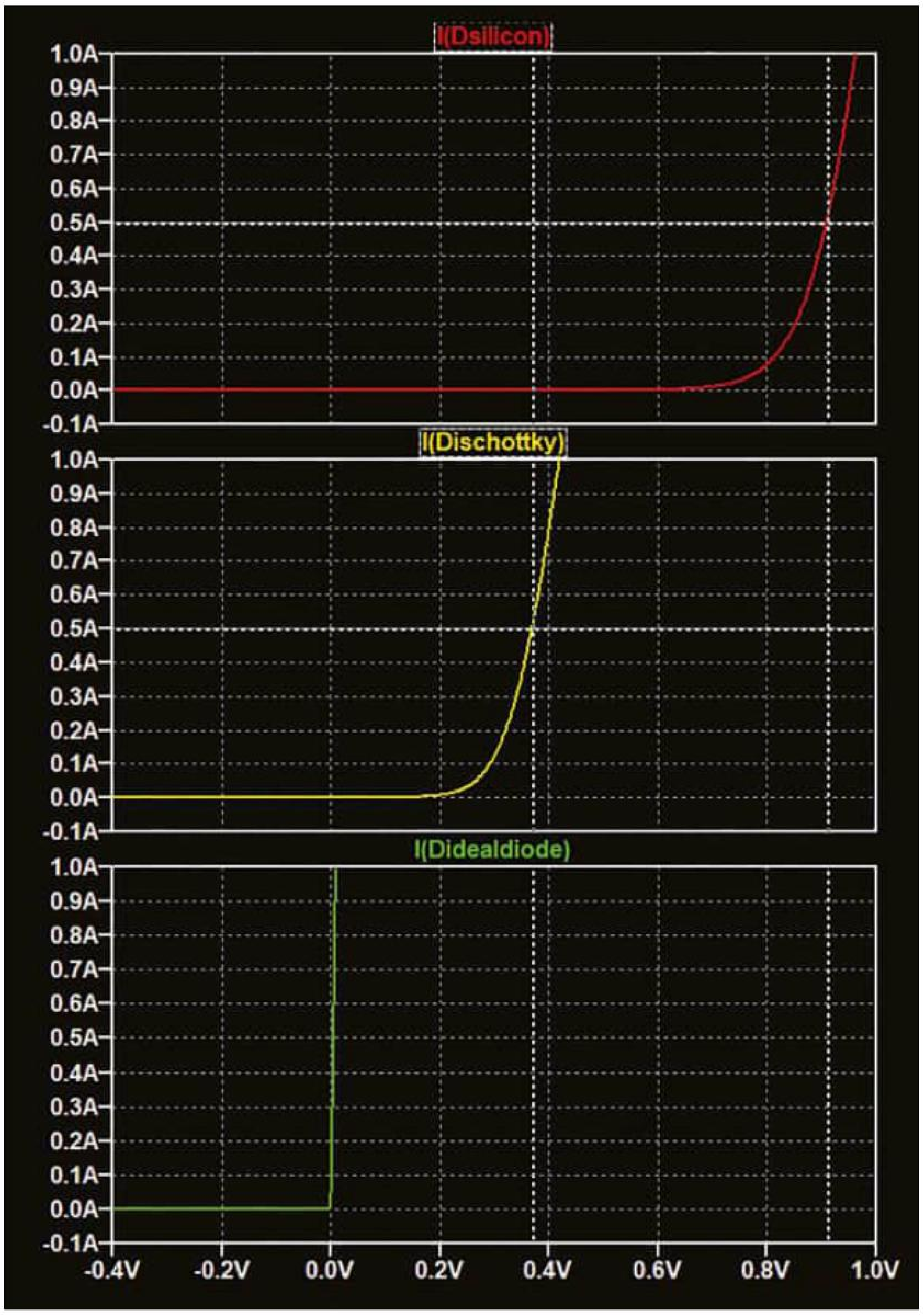


Fig.3. Results of the simulation shown in Fig.2.

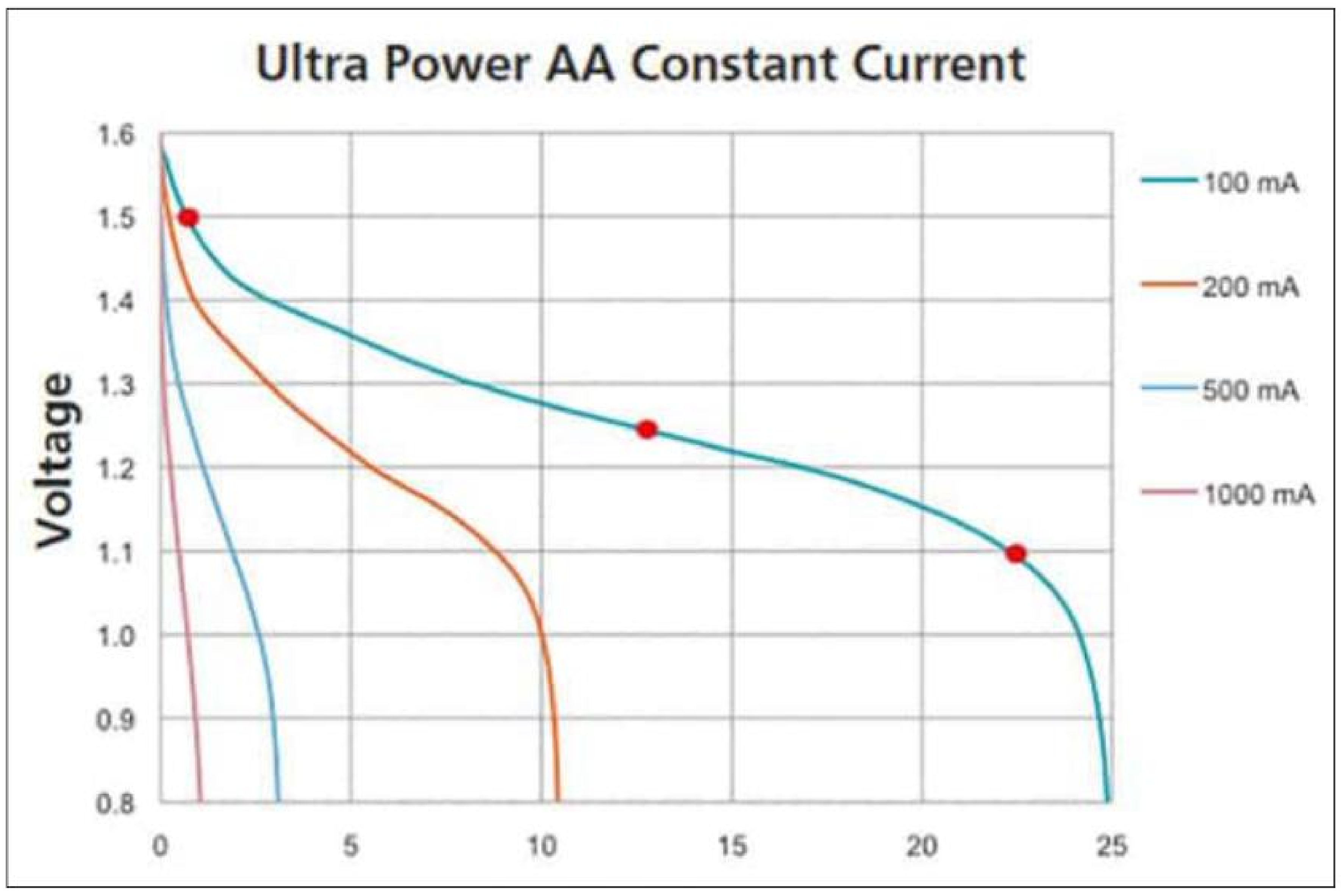


Fig.4. AA alkaline battery characteristics from Duracell datasheet.

which sweeps the voltage across all three diodes from –0.4V to 1.0V. Here we investigate the IV curves of the IN4001 silicon rectifier diode and the 1N5817 Schottky diode. The 1N5817 has a similar maximum forward current rating to the IN4001 (1A), but lower reverse voltage (20V). We also compare an idealised diode.

The 1N5817 is included in LTspice's library, but the 1N4001 is not, so we need to add a `.model` statement to the schematic (as shown in Fig.2). SPICE models for the 1N4001 are available online, including via the LTspice Wiki: **https://bit.ly/pe-jul23-ltsp**

The ideal diode is modelled with a forward voltage of 0V and very small on-resistance of 10mΩ. This simple piecewise model just switches between a resistance of 1MΩ below 0V and 10mΩ above 0V, but is sufficient to compare the shape of a close-to-ideal diode on the IV plots of the real diodes. In comparison, the models of the 1N4001 and 1N5817 are based on diode physics.

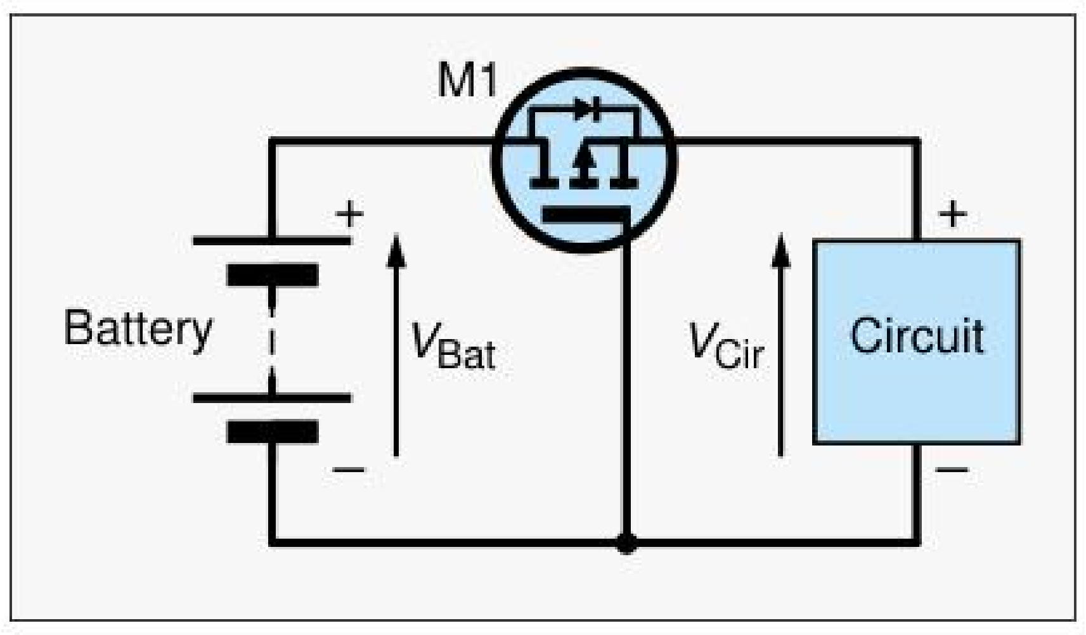


Fig.5. Battery-powered circuit with reverse-voltage protection using P-channel MOSFET.

Fig.3 shows the results from the simulation in Fig.2. The plots are configured to cover the same current range, up to 1.0A (the limit for the two real diodes). The plots show that, as expected, the Schottky diode forward voltage is lower than the silicon diode – at around 0.2-0.4V rather than 0.8-1.0V. This is slightly higher than the 'typical' values (eg, 0.6 to 0.7V for silicon diodes) due to the relatively high current. The ideal diode conducts above 0V, with the trace showing an almost vertical line due to the very low on-resistance.

For comparing diodes used for reverse power protection we can assume a certain supply current from the circuit and look at the diode voltage drops at this current. Fig.3 has cursors placed in the real diode traces at a current of 500mA. In LTspice these voltages can be read as 909mV for the 1N4001 and 369mV for the 1N5817 (these values can also be read more approximately from Fig.3). The ideal diode current can similarly be read using LTspice, but given that this is a linear piecewise model switching to R_{ON} at 0V, it is also given by $I_F R_{ON} = 500\text{mA} \times 0.01\Omega = 5\text{mV}$.

Using $V_F I_F$ we find the power dissipation to be 454mW for the silicon diode; 185mW for the Schottky diode; and 2.5mW for idealised diode. Using a Schottky diode more than halves the power wasted in the diode in this example. Of course, the power dissipated in the idealised diode is much lower. In addition to simply wasting power, designs that involve high dissipation in a reverse protection diode may require a heatsink for the diode, increasing cost, size and complexity. Choice of diodes (Schottky or silicon) should not only be determined by the continuous supply current for the circuit, but also the potentially large current that may flow at switch-on (inrush current), particularly if the circuit has large capacitors across the power supply that have to charge via the diode.

Battery considerations

We mentioned above how the use of a reverse protection diode can reduce effective battery life. Consider a rough example to illustrate this. Say we have a circuit operating from two AA 1.5V alkaline batteries. Assume the circuit will

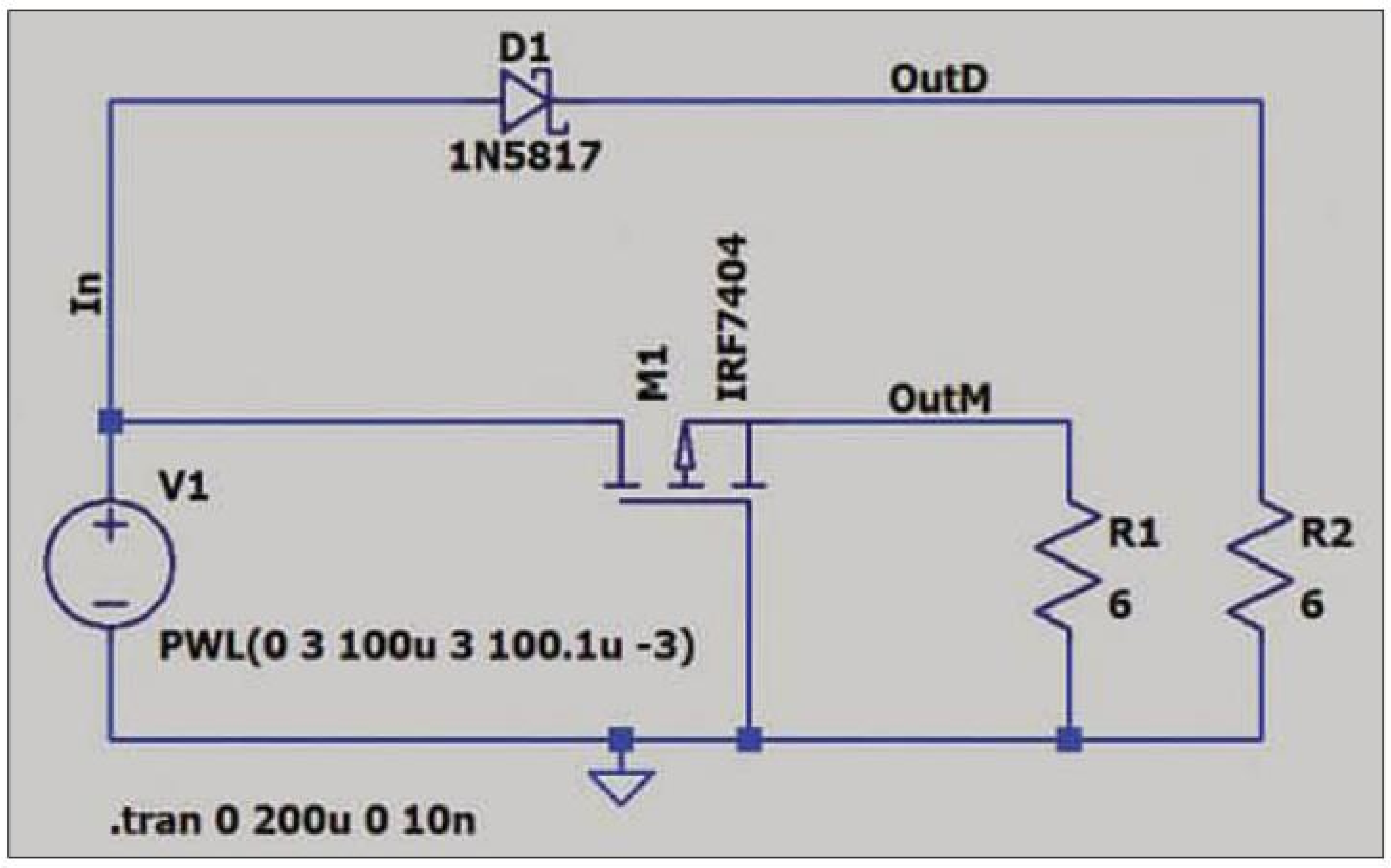


Fig.6. LTspice circuit to compare Schottky diode and MOSFET reverse-voltage protection

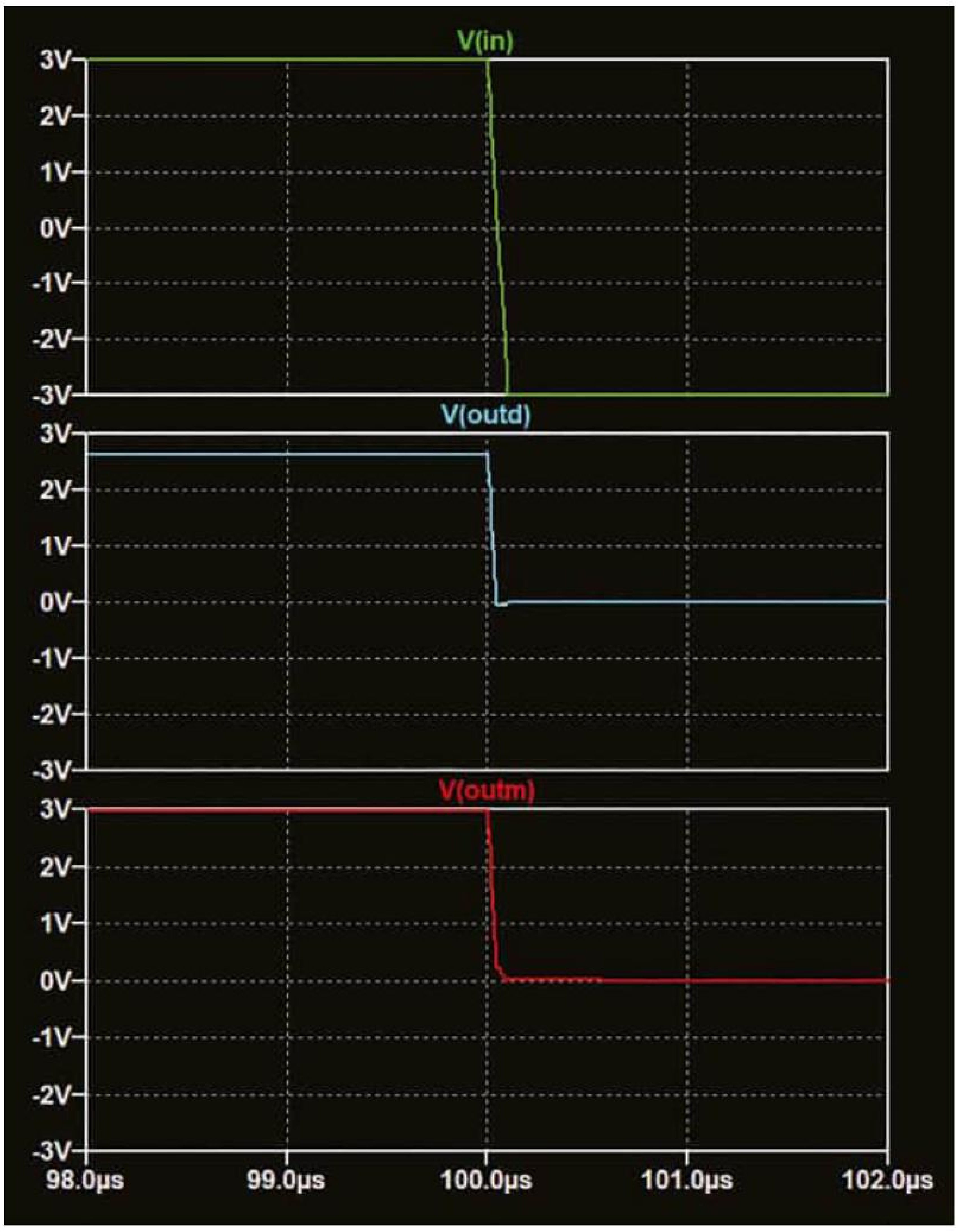


Fig.7. Results of the simulation shown in Fig.6.

work down to 2.2V and consumes 100mA. To understand what might happen we need to look at the discharge curve for the battery – these can be found on the manufacturer's datasheet. As an example, Fig.4 shows the constant-current discharge characteristics for a Duracell Ultra Power MX1500 1.5V AA (LR6) Alkaline-Manganese Dioxide Battery.

If we connect the batteries directly to the circuit, and assume that they both discharge exactly equally, then the circuit will function until the battery voltage reaches about 1.1V (total 2.2V), this occurs after about 23 hours and represents most of the available life of the batteries. If we use the 1N4001 as a reverse-protection diode the voltage drop is about 0.8V at 100mA, so each battery has to provide 1.1 + 0.8/2 = 1.5V for the circuit to function. This reduces the usable life to only 1 hour. With the 1N5817 Schottky diode the voltage drop is about 0.3V at 100mA, so each battery has to provide 1.1 + 0.3/2 = 1.25 V, which corresponds to a usable life of about 13 hours. The usable life points are shown by red marker dots on Fig.4.

The Schottky diode is considerably better than the silicon diode in this example, but still has a large impact on usable battery life. As we noted earlier, the Schottky diodes will also dissipate less power so are usually the preferred option compared with silicon diodes. Schottky diodes are more expensive (for example the 1N4001 is about £0.10 and the 1N5817 is about £0.30 for one-offs from a major UK distributor at the time of writing). These costs are unlikely to be an issue for constructing individual circuits but may be more important for larger-scale manufacture in some cases. Do bear in mind that a Schottky diode's higher reverse leakage may be important in some applications.

MOSFET protection

It is possible to use MOSFETs in place of diodes for reverse-voltage protection. Fig.5 shows a version of the circuit in Fig.1 with the diode replaced by a P-channel MOSFET. The diode shown in Fig.5 is the body diode of the MOSFET, not a separate component – however, it is important that it is in the same direction as the protection diode in Fig.1. The body diode will conduct initially when power is switched on, allowing source voltage (at the circuit's positive supply) to rise. Once the circuit voltage exceeds the MOSFET threshold voltage the transistor will switch on and conduct the current from the battery. Because this is a P-channel device the gate-source voltage must be negative with respect to the gate for conduction, which is what occurs in the situation shown in Fig.5. If the battery is reversed the gate-source voltage required for MOSFET conduction does not occur and supply current does not reach the circuit. A similar circuit with an N-channel MOSFET in the ground connection can also be used.

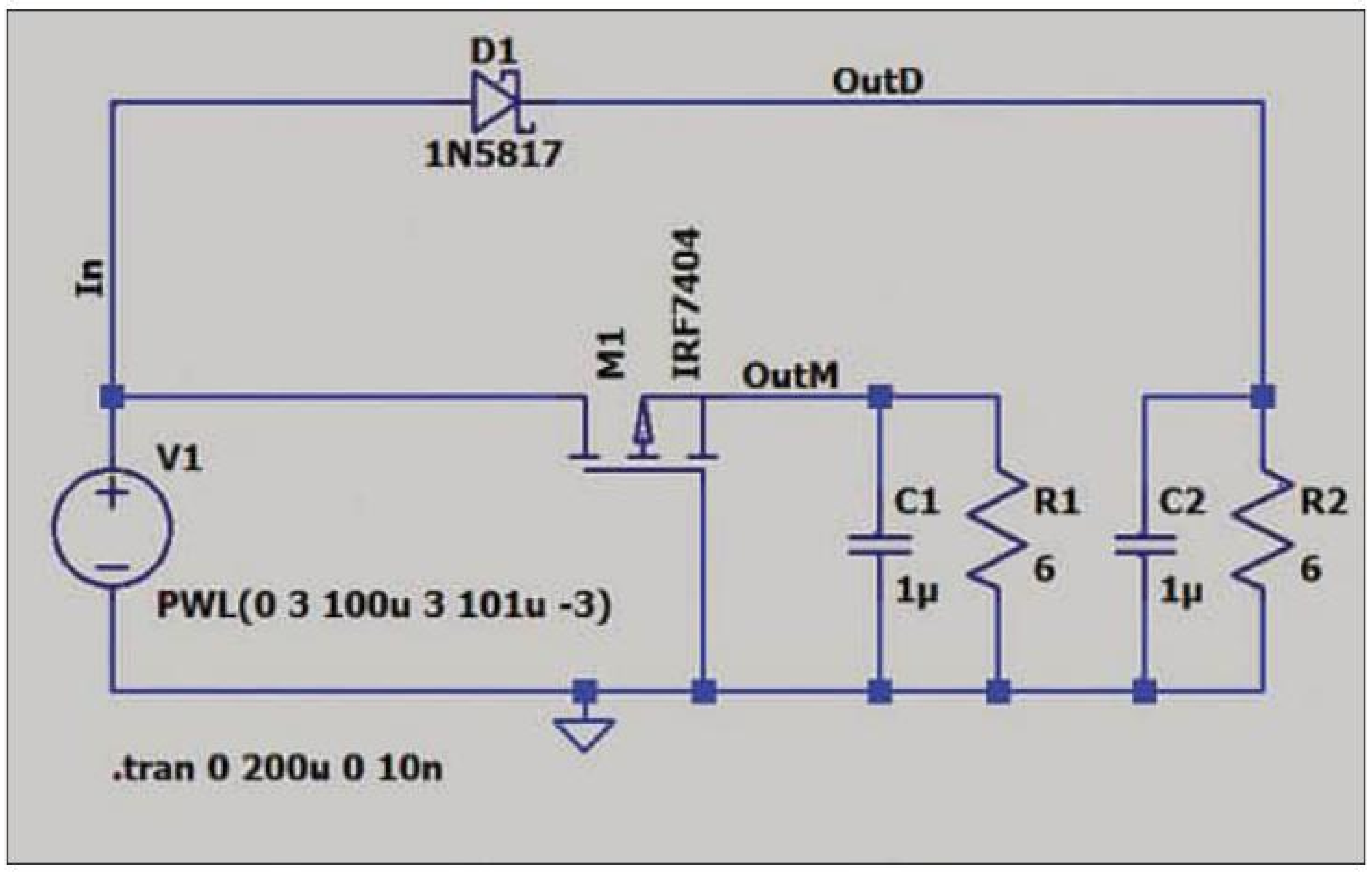


Fig.8. LTspice circuit from Fig.6 with capacitors across the circuit supply.

The circuit in Fig.5 has the advantage that the voltage across the MOSFET can be significantly smaller than a Schottky diode in the same application. The MOSFET on-resistance can be in the tens of milliohms range or lower, resulting in low dissipation. An LTspice circuit to simulate the arrangement in Fig.5 is shown in Fig.6. The 1N5817 Schottky is included for comparison. The circuit is represented by a 6Ω resistor for both devices. The MOSFET used is the IRF7404, which was a somewhat arbitrary pick from the LTspice library. It is rated at 6.7A, which is much higher than the 500mA flowing in this example. Simulation applies a 3V supply in the correct orientation for the first 100µs and then reverses it to –3V.

The results of the simulation in Fig.6 are shown in Fig.7. It can be seen that both the Schottky diode and MOSFET protect the circuit from the reverse-voltage (`OutD` and `OutM` are both zero after the input voltage reversal). While the correct polarity voltage is applied the forward-voltage drop across the Schottky diode of about 360mV is visible on the `OutD` trace. The `OutM` trace is very close to the 3V supply, with a measured drop of about 22mV. At 500mA this is a dissipation of 11mW, which is significantly smaller than for the 1N5817 at 500mA, as calculated above. The much lower voltage drop across the MOSFET would also lead to a close-to maximum available battery life in the scenario discussed above.

Fig.8 shows a version of Fig.6 with capacitors added across the circuit power supply. This is to be expected, as many circuits will have decoupling capacitors, and in scenarios where transient supply reversals or shorts could occur, larger capacitors may be used to hold up the supply to the circuit during the transient, or provide time for systems such as those using microcontrollers to such down cleanly if appropriate. As with the other examples, the devices and values used in Fig.8 are for illustrative purposes and do not represent recommendations or a specific design solution.

The results are shown in Fig.9. The voltages across the two circuits are shown in the top pane. It can be seen that the voltage across the circuit protected by the MOSFET (represented by the resistor `R1`) drops relatively quickly. This is because, unlike the diode, the MOSFET is able to conduct significantly in the reverse direction, returning energy from

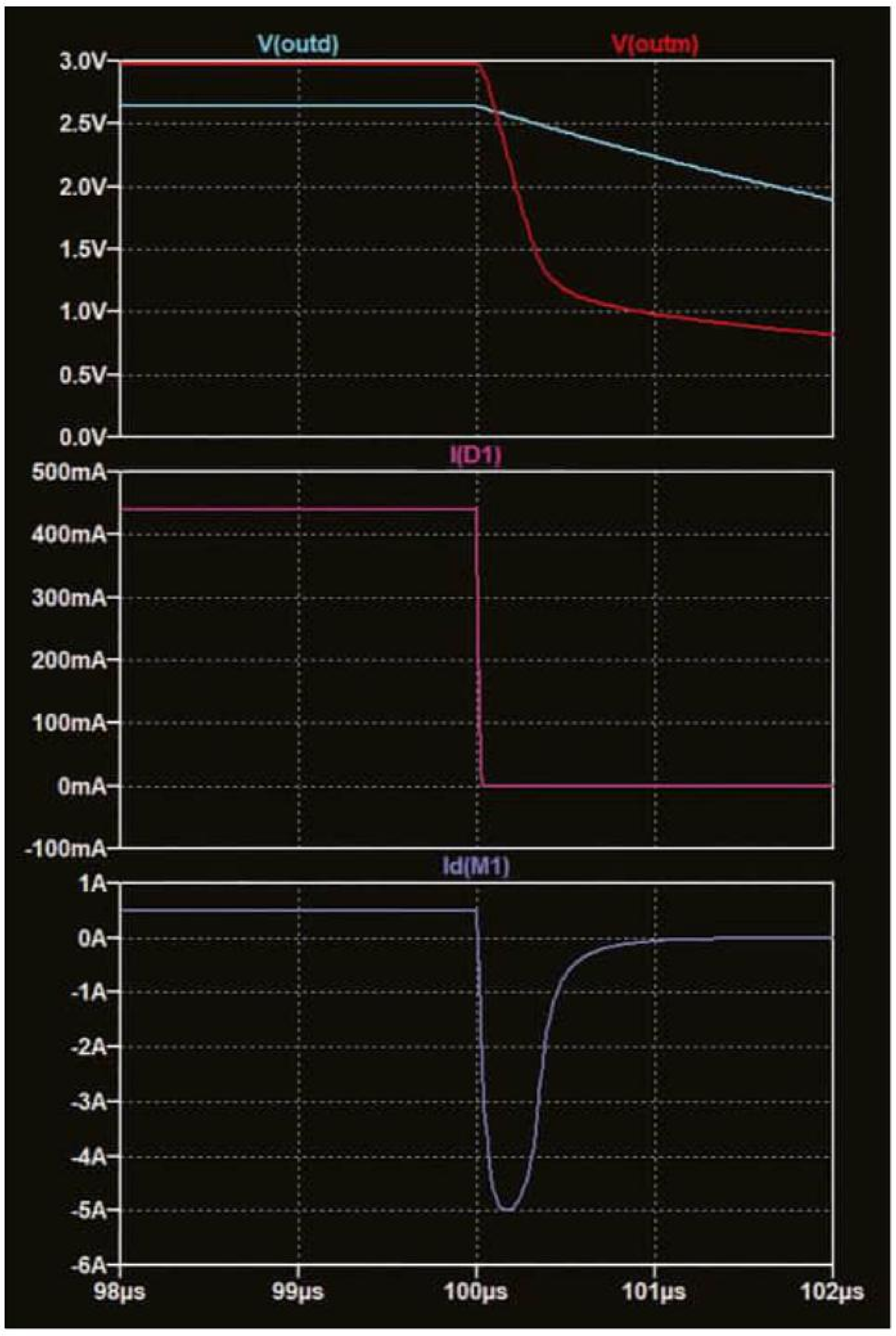


Fig.9. Results of the simulation shown in Fig.8.

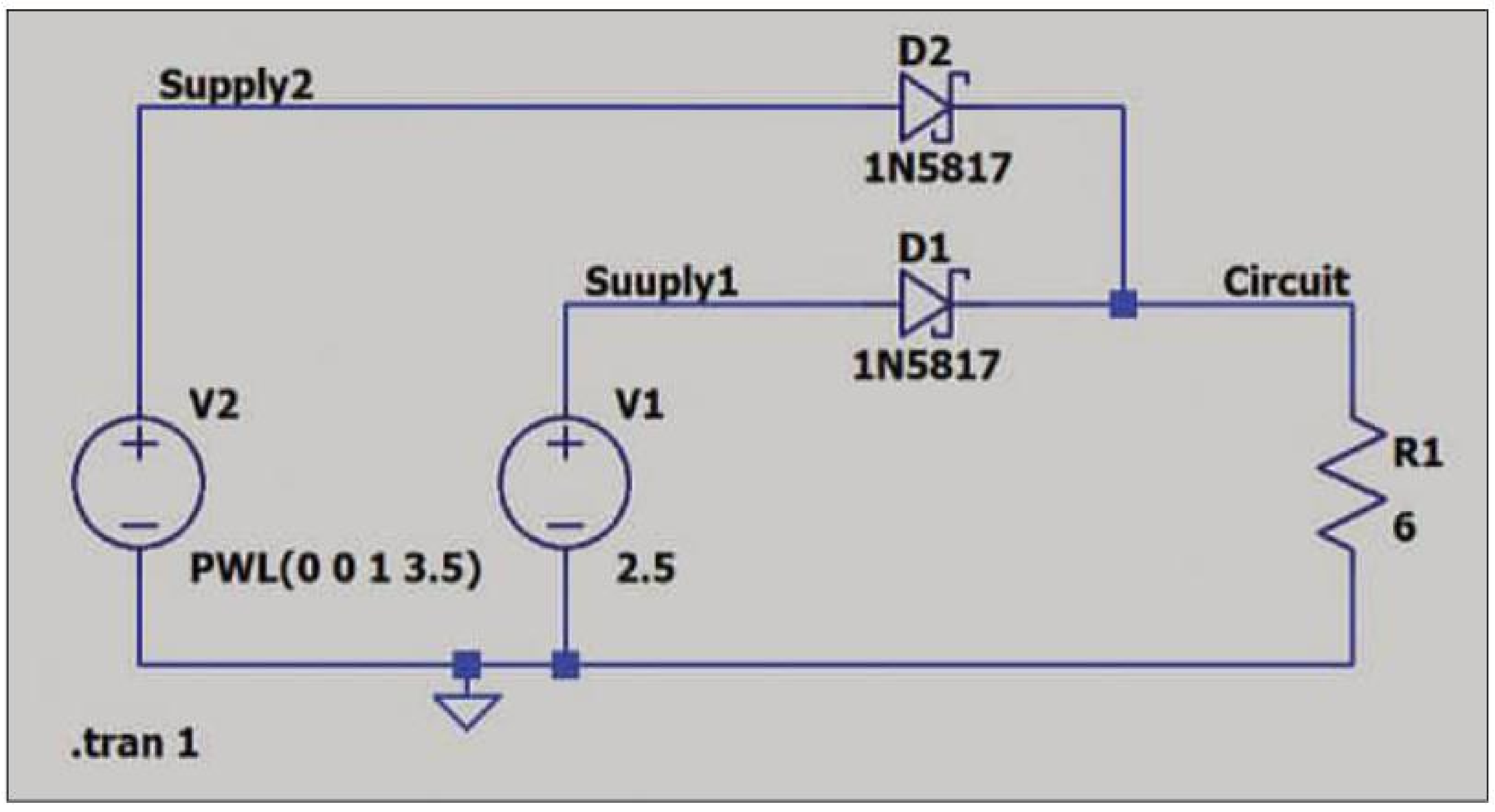


Fig.11. LTspice schematic to illustrate diode-OR power source circuits.

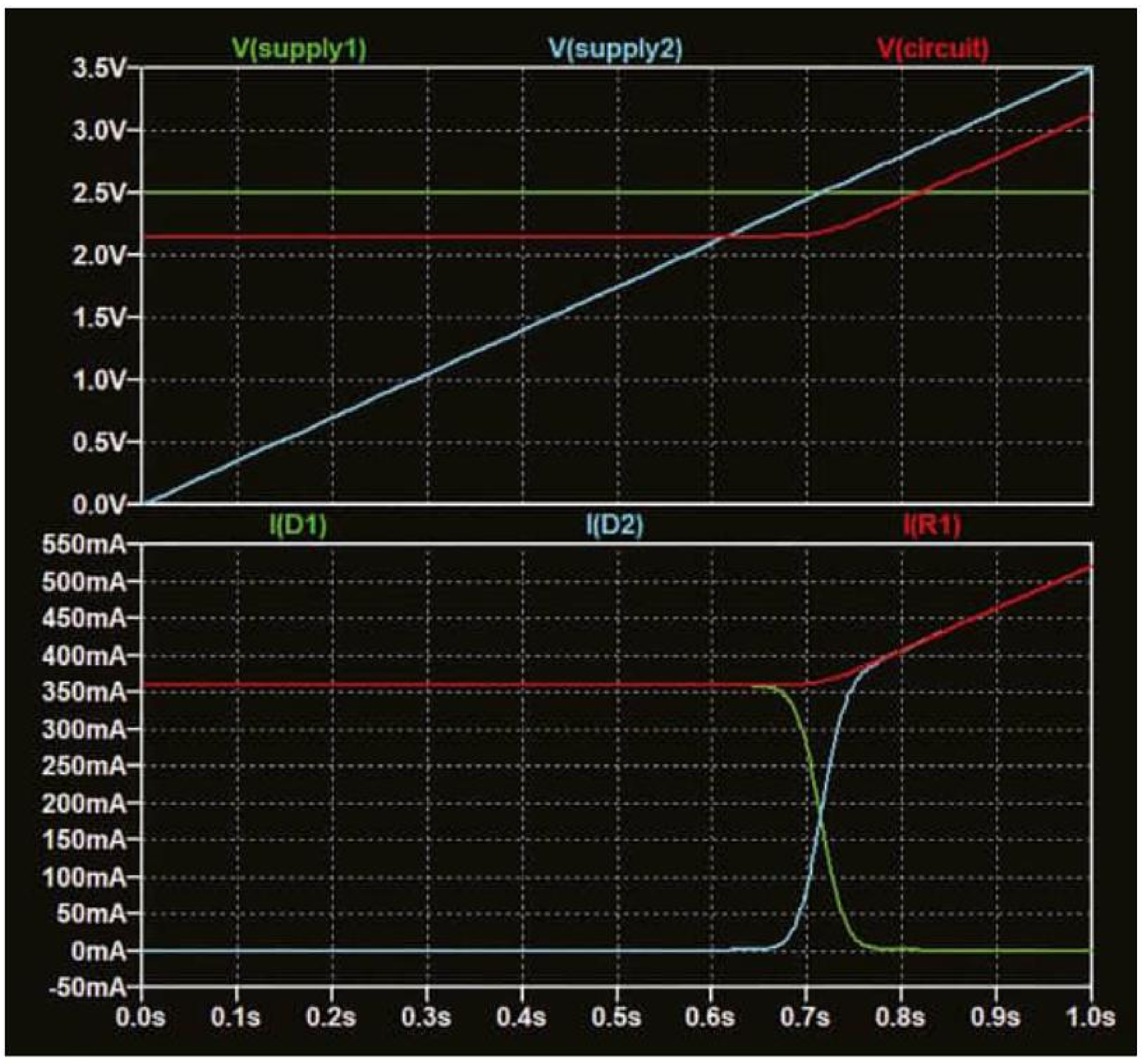


Fig.12. (right) Results from the simulation in Fig.11.

the capacitor to the battery – the reverse current pulse can be seen on the `I(M1)` waveform (MOSFET drain-source current). This lack of reverse-current protection from the MOSFET will be undesirable in some situations and significantly limits the possibilities of a capacitor being used to maintain the supply voltage during transient reversals. The diode has the disadvantage of higher voltage drop and power dissipation, but its reverse current is close to zero, so it protects against energy return to the battery, and the capacitor is better able to maintain the supply, although a small current will flow due to reverse leakage.

diode-OR

A circuit related to the reverse supply protection we have been discussing is the diode-OR circuit shown in Fig.10. There are two power supplies for the circuit. If only one supply is connected then its associated diode will conduct, and the circuit will behave in the same way as the reverse-protection circuits discussed above. If both supplies are present, with one sufficiently higher than the other, then the diode for the supply with the higher voltage will be forward biased, and the other reverse biased. The higher supply will provide power to the load. If both supplies have the same, or very similar voltage then both diodes will conduct and both supplies will deliver power to the load (this is called load sharing).

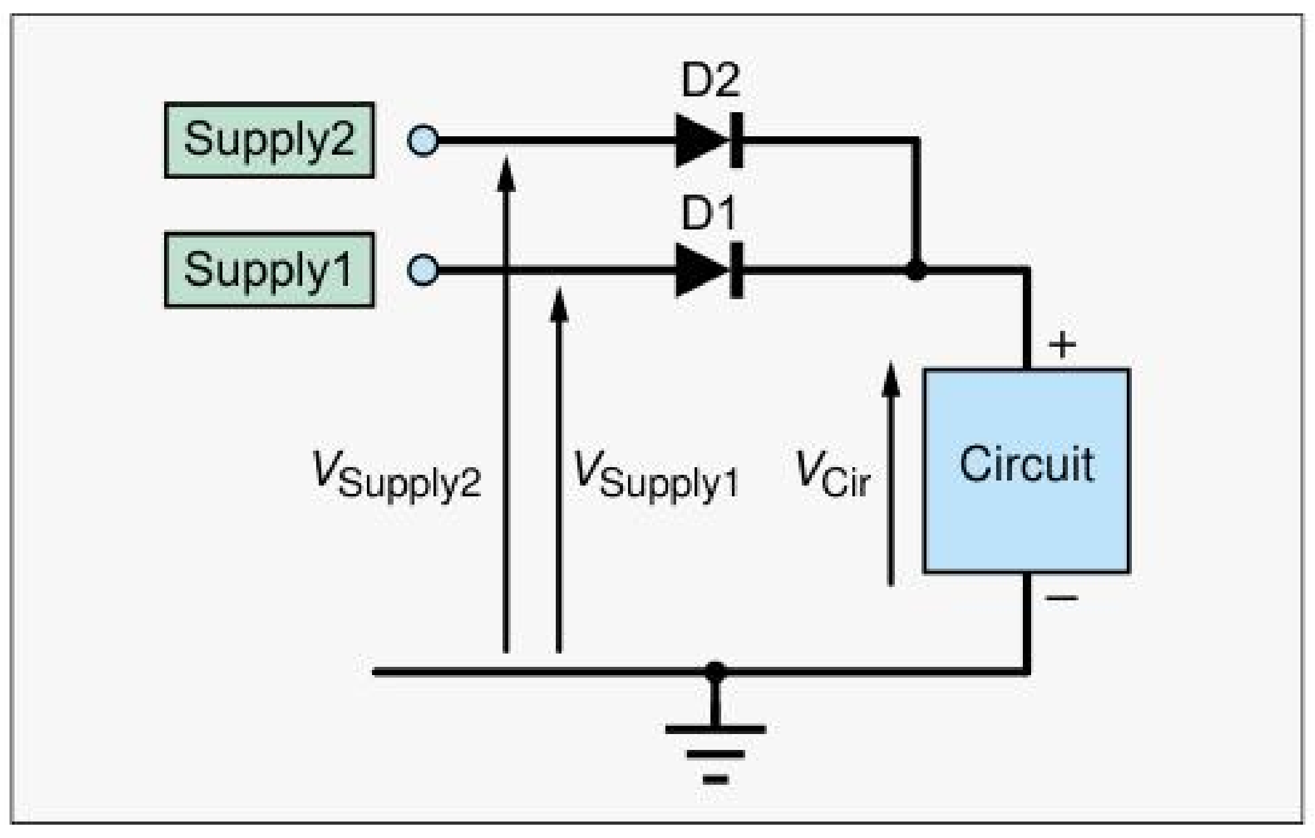


Fig.10. Diode-OR power source circuit.

The diode-OR circuit is not restricted to two inputs, more supplies can be connected via diodes. This type of circuit has a number of applications, including providing a means for mains DC power adaptors to take over from internal batteries when they are plugged into a product, or switching from primary to backup batteries. Load sharing can in principle be implemented as noted, but may be difficult to achieve effectively in practice due to differences in individual diodes and thermal effects.

Fig.11 shows an LTspice circuit to illustrate the principles of diode-OR power source circuits. `Supply1 (V1)` is steady a 2.5V while `Supply2 (V2)` ramps from 0V to 3.5V over one second. The resistor `R1` is used to represent the circuit being powered. Schottky diodes are used for the reasons discussed above. The results are shown in Fig.12. The upper plane shows the voltages. The circuit voltage remains at about 2.15V (2.5V from `Supply1` minus the 350mV forward voltage drop of `D1`) until the voltage on Supply2 gets close to 2.5 at around 0.7s. The circuit voltage then increases, following the increase in `Supply2`, staying at approximately 350mV below `Supply2` due to the diode voltage drop. The lower pane in Fig.12 shows the currents. From 0.68 to 0.76s the two diodes both supply current to the circuit – load sharing is occurring. At around 0.714s `Supply2` equals `Supply1` at 2.5V and the diode currents are equal.

Our discussion on reverse-voltage protection and diode-OR circuits has shown the usefulness of diodes in power supply

Simulation files

Most, but not every month, LTSpice is used to support descriptions and analysis in *Circuit Surgery*.
The examples and files are available for download from the *PE* website.

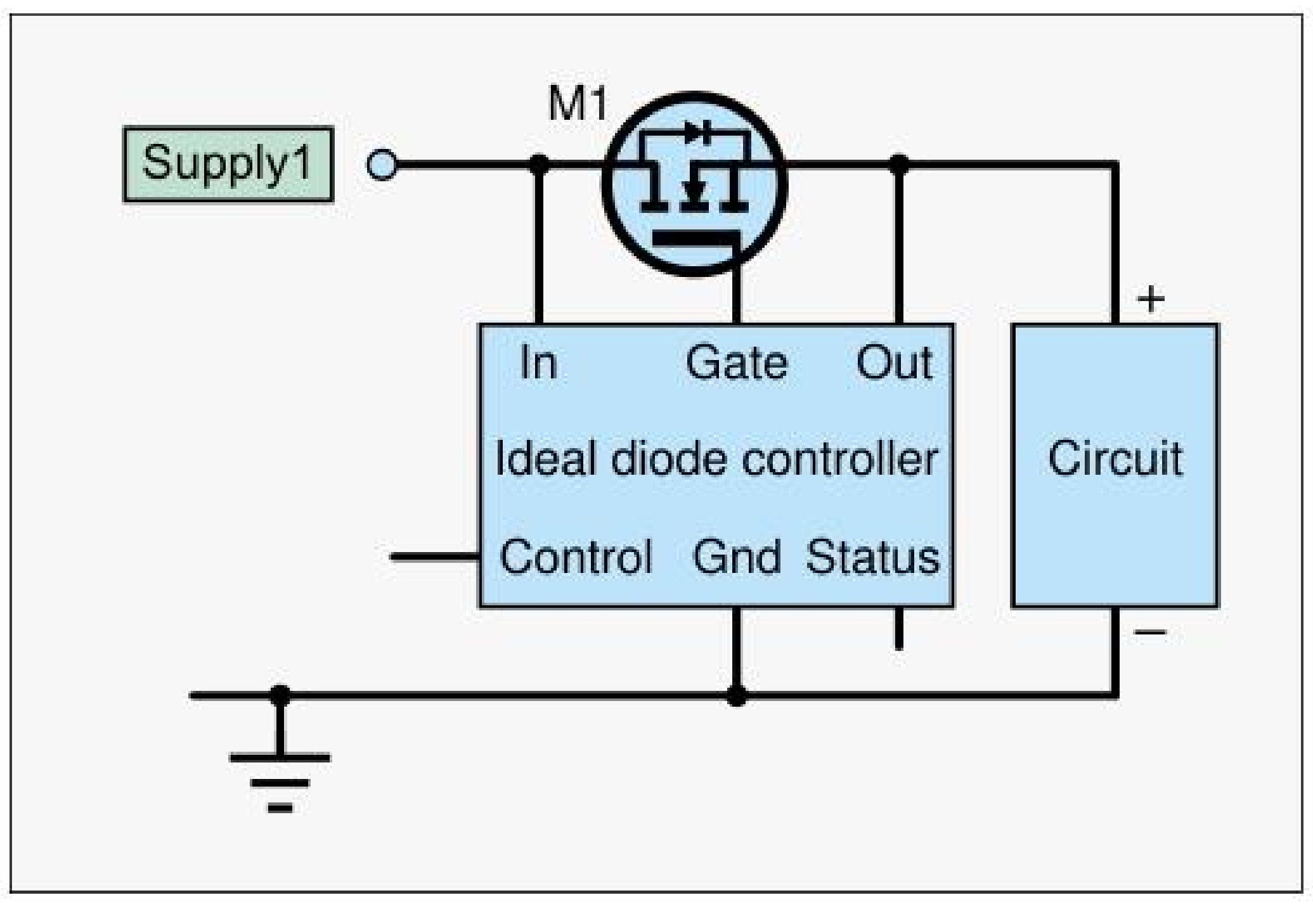


Fig.13. Ideal diode controller concept circuit (reverse protection).

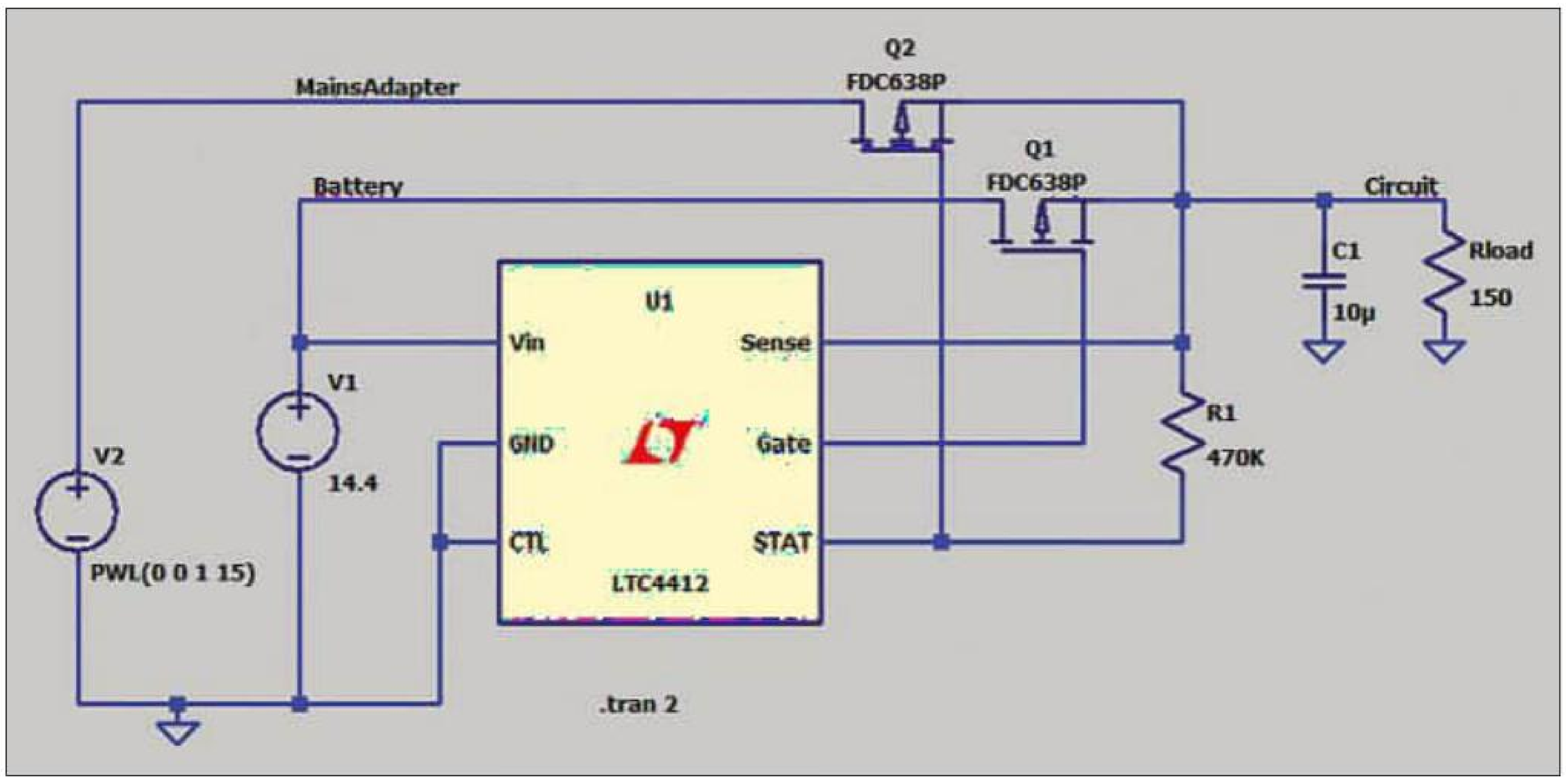


Fig.15. LTspice circuit for LTC4412 used to control switchover between a battery and a mains adapter (based on example file from Analogue Devices).

circuits, but has also highlighted some deficiencies in terms of power dissipation and the impact on supply voltage and hence battery life in battery-powered systems. We have also seen that MOSFETs can be used in place of diodes to significantly reduce voltage drop and power dissipation. However, MOSFETs may conduct in the reverse direction in some cases, which can cause significant problems in some applications. Furthermore, the basic MOSFET circuit in Fig.5 may not work in all cases – for example, if the supply voltage is not correctly matched to the required gate-source voltage. For high voltages, a potential divider can be used to reduce the gate voltage, but if the supply voltage is too low the solution is more complex – for example, a charge pump can be used to obtain the voltage required to drive the gate. The diode-OR circuit could also make use of MOSFETs, but providing the gate control is not as straightforward as for simple reverse supply blocking.

Improved protection designs

All of the issues above can be addressed by using MOSFETs in place of the diodes with more sophisticated control of the MOSFET switching, along with other functionality such as charge pumps for gate drive voltages (if needed). There are a large range of integrated circuits from various manufacturers which provide this general capability, with different specific features and target applications. These chips are often referred to as 'Ideal Diode Controllers' because they control MOSFETs to perform like ideal diodes, although some devices are given more specific names such as 'Reverse Polarity Protection Controller' and 'Diode-OR Controller'.

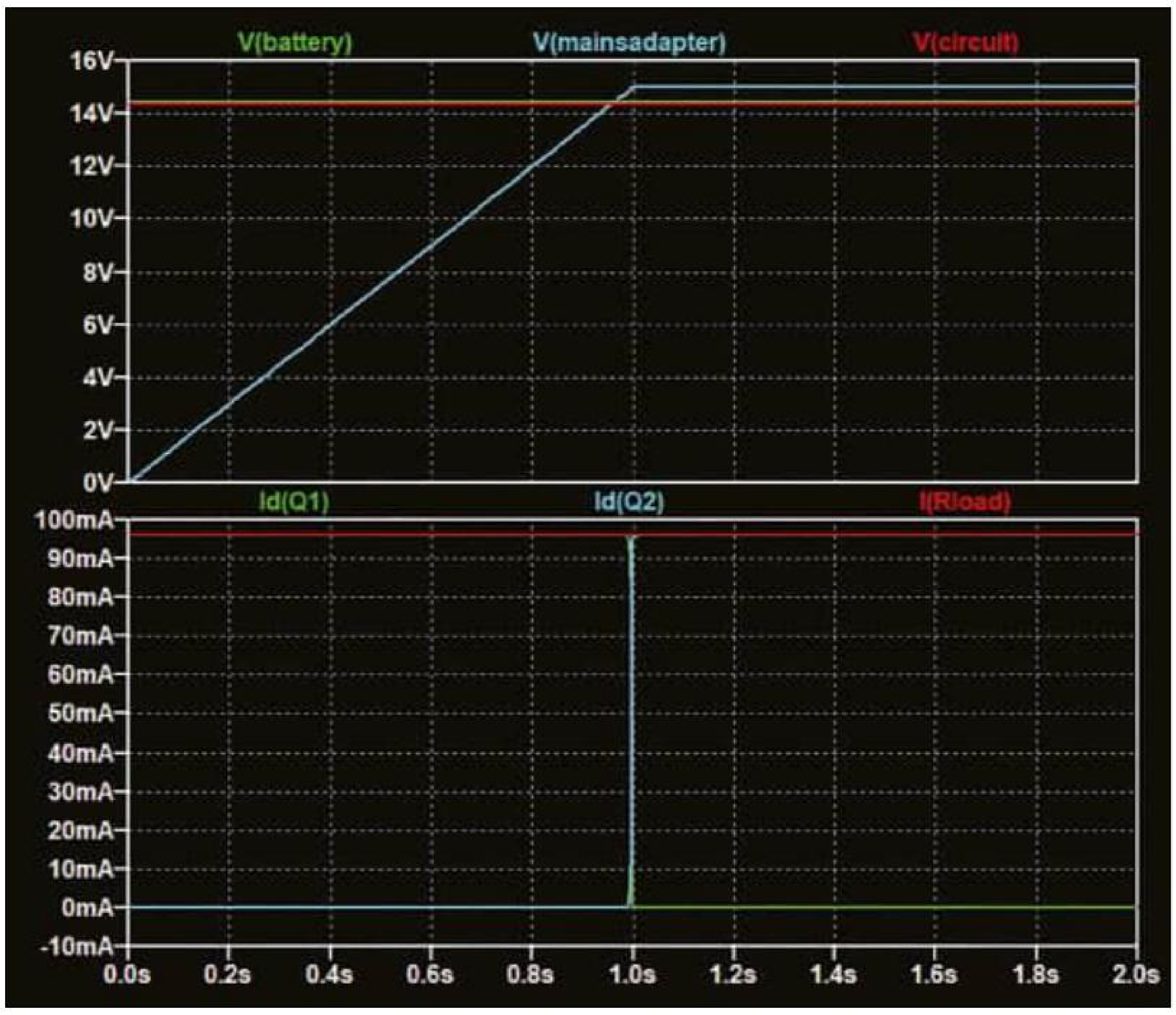


Fig.16. Results from the simulation in Fig.15.

Fig.13 and Fig.14 show concepts of how ideal diode controllers are wired for reverse protection and diode-ORing. These are generic examples and do not represent specific devices, which vary, so individual datasheets must be consulted for connection and operational details. The circuits in Fig.13 and Fig.14 use external N-channel MOSFETs, but some ideal diode controllers use P-channel devices, and some have internal MOSFETs. Compared to Fig.5, Fig.6 and Fig.8, the MOSFETs in Fig.13 and Fig.14 have their source and drain reversed so that the body diode is in the same direction as the P-channel device used above. This is not a problem because the MOSFETs can conduct in either direction.

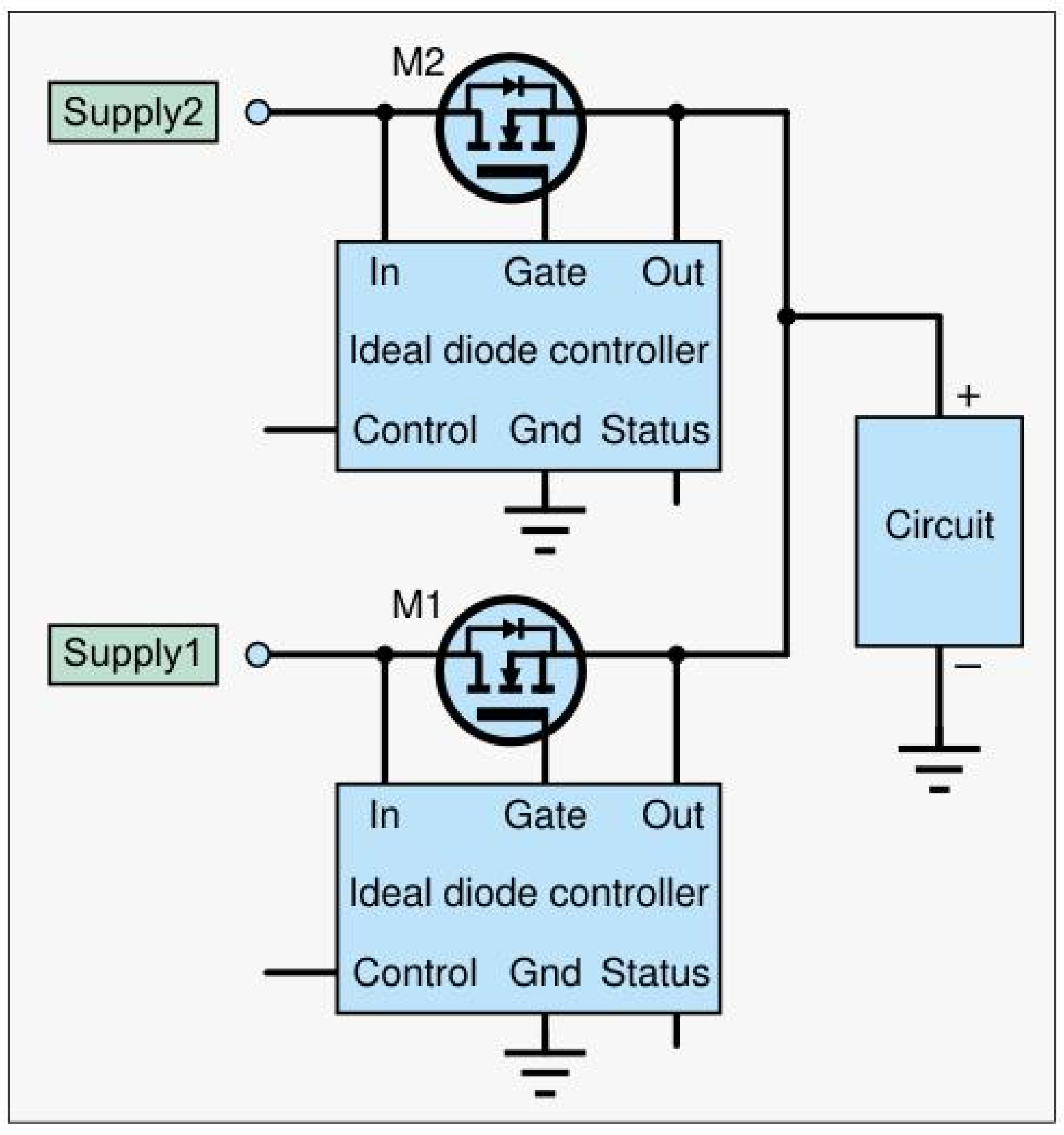


Fig.14. Ideal diode controller concept circuit (diode-OR).

Ideal diode controllers have pins connected to the input supply and the circuit being powered that allows them to sense what is happening and hence control the MOSFET gate appropriately. Their own power may be a separate pin, but it will often be connected to either the input or circuit supply. Some ideal diode controllers provide status and/or fault condition outputs, which can be read by a microcontroller. Some have control inputs, such as 'enable' to provide low-power shutdown mode. In load sharing applications ideal diode controllers will regulate the MOSFET gate-source voltage to provide stable operation.

Fig.15 shows an LTspice schematic for an example circuit using an ideal diode controller. This uses the LTC4412 'Low Loss PowerPath Controller' from Linear Technology (now from Analogue Devices) to implement switchover between a battery and a mains power adapter. A number of ideal diode controller ICs are available in the LTspice library. The circuit in Fig.15 is based on Fig.2 in the LTC4412 datasheet and is adapted from an LTspice file available from Analogue Devices. The simulation is similar to that in Fig.11 and Fig.12. The mains adapter voltage (from `V2`) is ramped from 0 to 15V (then stays at 15V) while the battery voltage remains fixed at 14.4V throughout. When the sense pin voltage exceeds the V_{in} pin by 20mV `Q1` is switched off and the STAT pin goes low, turning on `Q2` and allowing the mains adapter to power the circuit. The results in Fig.16 show that the circuit voltage remains almost constant, but switchover can be seen in the current waveforms for the MOSFETs in the lower pane.

Max's Cool Beans

By Max the Magnificent

Arduino Bootcamp – Part 7

I can't believe we are already slip-sliding our way into the second half of the year. Sad to relate, I haven't prepared a speech and I don't have anything appropriate to wear. When I was a kid, it seemed elderly folks liked nothing better than to endlessly babble about how time seems to pass faster the older you get. You can only imagine my surprise to discover that the ancients actually had a clue as to what they were droning on about. Now, when I blink, another day disappears. When I sneeze, I lose another week. And if I… but we digress.

Round, square, curly…

I really enjoy receiving emails from members of the *PE* community, especially communiques that commence by conveying how wonderful I am (hint, hint). By some strange quirk of fate, I recently received such a missive from someone we'll call Alan (because that's his name). After making mention of how much he was enjoying our *Arduino Bootcamp* series (well played Alan), he continued as follows: 'Whilst I have done some programming in the past, including BBC BASIC and assembly language for use on PIC micros, C/C++ is new to me, so I appreciate the way in which your series explains how the programs are built up. Having said this, I do have difficulty with some of the syntax, especially when to use curly brackets.'

'Well, strike me with a kipper,' I thought (I think this a lot, but don't ask me why), 'if there's one topic I love to waffle about endlessly, that topic is brackets in general and curly brackets in particular' (I really do need to get out more). As you are doubtless aware, there are four primary types of brackets (Fig.1). In British usage they are known as: 'round brackets' (or simply 'brackets'), 'square brackets', 'curly brackets' (known to some as 'squiggly brackets'), and 'angle brackets'; in American usage these are respectively known as: 'parentheses', 'brackets', 'braces', and 'chevrons'.

( )	[ ]	{ }	< >
Round brackets *or* parentheses	Square brackets *or* brackets	Curly (squiggly) brackets *or* braces	Angle brackets *or* chevrons

Fig.1. Different types of brackets.

Different programming languages make use of some or all of these symbols for diverse purposes. In fact, the same language can employ the same symbols in multiple roles. Don't panic! This is nothing we're not used to. Consider the artful apostrophe, for example, which can serve to indicate possessive cases (Harry's hat), contractions (Harry's lost his hat), and omitted letters ('Arry's 'at) in regular written English.

Squiggly wiggly

As we previously discussed in a couple of *Coding Tips and Tricks* columns (*PE*, April and June 2020), in the case of C/C++, one use of curly (squiggly) braces is to denote a block of code in which variables can be declared. The most obvious example of this is as part of a function declaration and definition, for example:

```
void MyFunction () // Declaration
{
  // Zero or more    Definition
  // statements
}
```

Remember that, unlike some languages, C and C++ really don't care overmuch about whitespace characters such as the space, tab (ie, horizontal tab), vertical tab, line feed (newline), carriage return and form feed. Having said this, my personal preference is for { } pairs to be vertically aligned, as shown above, because this makes our code easier to parse and debug.

Also, if I'm using a simple editor, such as the one found in Version 1 of the Arduino's integrated development environment (IDE), I prefer to use 4-space indentation. In the case of a more sophisticated editor, such as the one found in Arduino IDE V2 (the version we decided we were going to use at the beginning of this series), which automatically indicates indented blocks of code, I use 2-space indentation to match what the editor wants to do.

Observe the comment in the above code saying, `Zero or more statements`. Why would we wish to create a function with nothing in it? Well, as we noted in an earlier column (*PE*, April 2023), one way to approach creating a new program is to start off with a skeleton (bare bones) framework, and this can include using one or more stub (empty) functions whose purpose is to check that they can be called from elsewhere in the program and that everything hangs together. We can return to fill in their contents later.

When we talk about the 'scope' of a variable, we are referring to the extent that this variable can be 'seen' by various portions of the program. For example, suppose we declare two integer variables as illustrated below:

```
int MyGlobalInt;

void MyFunction ()
{
  int myLocalInt;
  // More statements
}
```

In this case, `MyGlobalInt` is said to be a 'global variable' because it's declared at the top level of the program outside of any functions. This means that its scope is the entire program, and it can be seen and modified by any of the functions forming the program.

By comparison, `myLocalInt` is said to be a 'local variable' because it's declared inside `MyFunction()`. In this case, its scope is localised to `MyFunction()` and it cannot be seen or modified by any other functions. It also means that this variable is instantiated (brought into existence) when the function is called, and it 'evaporates into the ether' when the function terminates and returns control to whatever called it. This explains how we can declare variables of the same types and/or names in multiple functions without them being related to each other in any respect.

As a little reminder, I use the camel case naming convention for my function and variable names. With this convention, which is also known as Pascal case, multiple words are joined together without spaces. When it comes to global variables and function names, I use UpperCamelCase in which the first letter of every word is capitalised. By comparison, with local variables, I use lowerCamelCase in which the first word is all lowercase, including the first letter. Using this convention makes it easy to differentiate between global and local entities when reading and writing code.

Building blocks

If you wish, you can use pairs of curly brackets inside a function to gather groups of statements together to make it clear to yourself and anyone else that you consider these statements to be related. These are often referred to as 'blocks,' and using this technique may be referred to as 'block programming.'

Of particular interest to us here is that variables can be declared within a block, in which case the scope of those variables is limited to the block in which they are declared. Consider the following code snippet, for example:

```
void MyFunction ()
{
  int myLocalIntA;
  // More statements

  { // Start of block
    int myLocalIntB;
    // More statements
  } // End of block
}
```

In this case, the scope of `myLocalIntA` is the entire function, which means it can also be seen and modified from within our block. By comparison, the scope of `myLocalIntB` is limited to the block within which it is declared.

Furthermore, this also means that `myLocalIntB` is instantiated (brought into existence) when it is declared at the beginning of the block and it disappears at the end of the block. A little thought brings us to the realisation that we can have multiple blocks in the same function and (if we so desire) we can declare variables of the same type and/or name in two or more of these blocks, where these variables will be totally distinct from each other.

As one final point, we can have nested curly brackets, for example `{ { } }`, `{ { { } } }`, `{ { } { } { } }` and so on, which means we can have blocks within blocks *ad infinitum*.

Making a statement

Computer programs are made up of a series of statements. Each statement instructs the computer to perform a specific action. Different programming languages do things in different ways. In C/C++, a statement is terminated with a semicolon ';' character.

Let's assume we've already created a function, as part of which we've declared three local integer variables called `a`, `b`, and `c`. Somewhere in the function we might have a series of three statements as follows:

```
a = 6;
b = 4;
c = a + b;
```

Observe how each of these statements is terminated by a semicolon. Now suppose we decide to gather these statements together in a block as shown below:

```
{
  a = 6;
  b = 4;
  c = a + b;
}
```

In this case, observe that we still need to terminate each of the individual statements with semicolons, but we don't need (cannot use) a semicolon following the '}' that terminates the block. Similarly, we don't need (cannot use) a semicolon following the '}' that terminates the function itself.

The word 'compound' is defined as something that is composed of two or more separate elements. This explains why another name for blocks is 'compound statements' because they are composed of two or more simple statements.

There's a condition

We introduced the concept of `if()` conditional statements earlier in this series (*PE*, March and April 2023). Our newfound knowledge regarding curly brackets and compound statements (blocks) should serve to clarify our previous discussions. For example, assuming we've declared some local integer variables called `a`, `b`, `c`, `d` and `e`, we could have an `if()` associated with a single statement as follows:

```
if (a > b) c = 10;
```

In English this equates to 'If `a` is greater than `b` then make `c` equal to 10.' As we noted earlier, the C/C++ languages don't care overmuch about whitespace characters, so some programmers would write this statement as follows:

```
if (a > b)
  c = 10;
```

The main thing to note is that, in both of the above cases, the entire statement is terminated by a semicolon. Alternatively, we could associate a compound statement with our condition as follows:

```
if (a > b)
{
  c = 6;
  d = 4;
  e = c + d;
}
```

As usual, each of the simple statements is terminated with a semicolon, but don't need (cannot use) a semicolon following the '}' that terminates the compound statement.

It's also worth reminding ourselves that, even if we have only a single statement associated with an `if()` condition, my personal preference is to still employ a `{ }` block as illustrated below:

```
if (a > b)
{
  c = 10;
}
```

Although this may appear to be a case of 'overkill,' it makes things a whole lot clearer when you have multiple levels of `if()` … `else if()` … `else()` conditions. Also, it makes one's life a whole lot simpler when adding (and later removing) additional temporary statements while debugging recalcitrant code.

An array of confusion

I feel confident that things would remain relatively clear if we could terminate our discussions of curly brackets at this point. Unfortunately, these little scamps are used for additional purposes, such as the initialisation of arrays (also the initialisation of `enum` and `struct` constructs, which we have not yet introduced).

We said hello to arrays in an earlier column (*PE*, February 2023). Suppose we declare a 3-element integer array called `x`, as illustrated below:

```
int x[3];
```

Observe the use of the square brackets to (a) indicate that this is an array and (b) specify the size of the array. Remember that we always start counting from zero, so the elements of this array will be numbered (indexed) as 0, 1, and 2. Later in the program, we might load values into our array as follows (these values have no meaning beyond serving as an example):

```
x[0] = 7;
x[1] = 2;
x[2] = 5;
```

In this case, we are using square brackets to indicate elements of interest. Alternatively, we can use curly brackets to initialise our array as part of its declaration as illustrated below:

```
int x[3] = {7, 2, 5};
```

There are several points to observe here. First, the initialisation values are comma-separated. Second, no comma is required (or allowed) after the final value. Third, since this entire construct is a statement, it's terminated with a semicolon, which appears outside the closing '}' character.

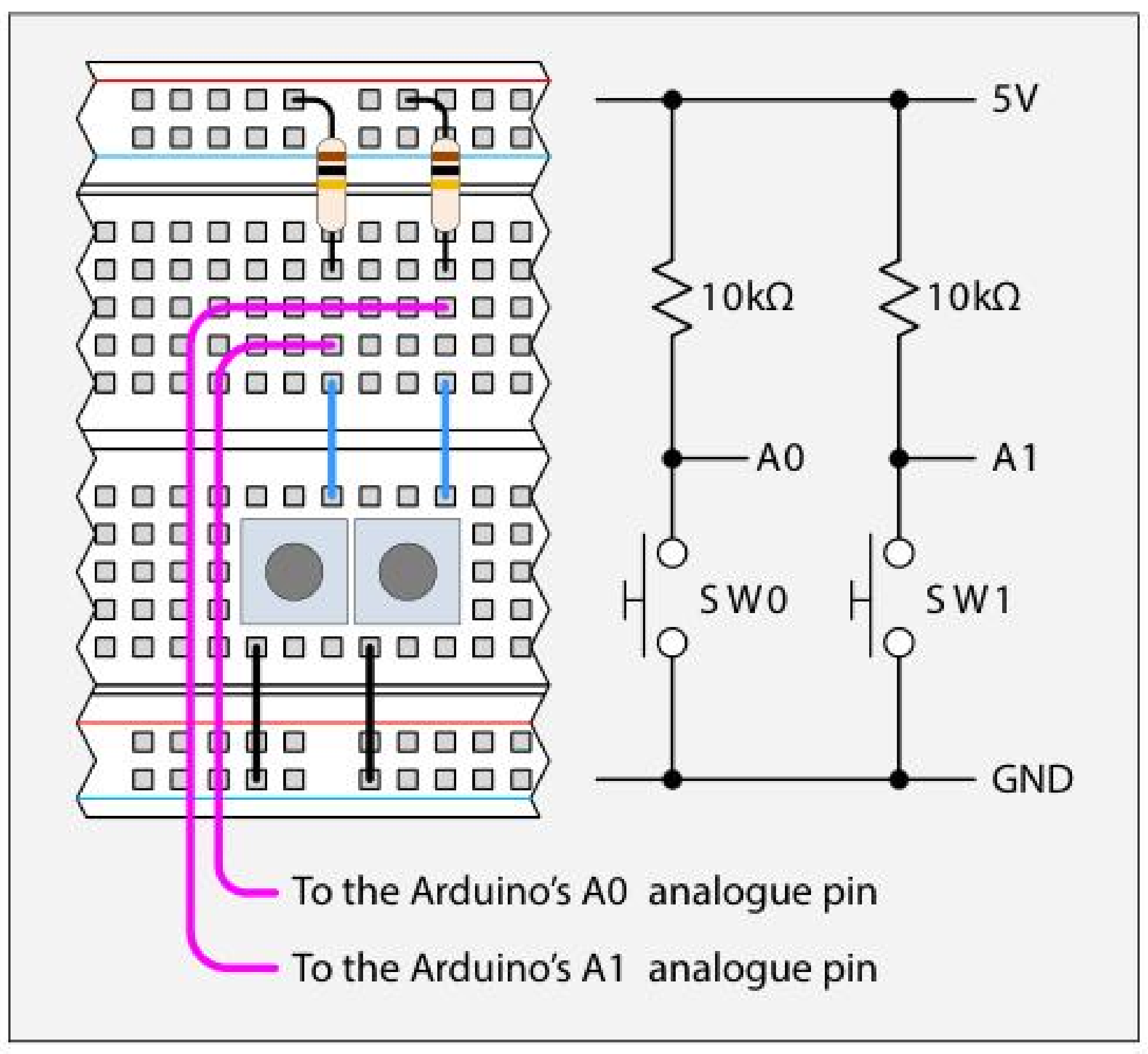

5V
10kΩ
10kΩ
A0
A1
SW0
SW1
GND
To the Arduino's A0 analogue pin
To the Arduino's A1 analogue pin

Fig.2. Adding a second pushbutton switch.

The fact that a semicolon is required here but not to terminate a compound statement can be endlessly confusing to beginners, but it all makes sense when you think about it. Trust me, have I ever lied to you before? You don't know? Well, that just goes to show how good I am.

As you may recall, we've seen something like this before when we first declared and initialised the array defining the segments that correspond to the numerical characters 0 to 9 on our 7-segment display (*PE*, April 2023):

```
byte DigitSegs[NUM_DIGITS] =
     {
        // ABCDEFGDP
           B11111100, // 0
           B01100000, // 1
           B11011010, // 2
           B11110010, // 3
           B01100110, // 4
           B10110110, // 5
           B10111110, // 6
           B11100000, // 7
           B11111110, // 8
           B11110110  // 9
     };
```

Note the lack of a comma after the final value and the presence of a semicolon after the closing '}' character. Now everything starts to make more sense, doesn't it?

Multidimensional madness

While we are here, we may as well consider the concept of multidimensional arrays in C/C++. The simplest form is a two-dimensional (2D) array, but 3D, 4D, 5D, and higher dimensions may also be employed as required. For example, supposing we wish to declare and initialise a 2D integer array called x with three rows, each comprising five columns, we could do so as follows (again, these values have no meaning beyond serving as an example):

```
int x[3][5] =
{
  {10, 12, 99, 42, 45},
  {18, 15, 21, 36, 54},
  {72, 91, 14, 62, 77}
};
```

In this case, our three rows are numbered 0, 1, and 2, while our five columns are numbered 0, 1, 2, 3, and 4. If we subsequently wished to change the value of the fourth element in the second row (indicated in red) from 36 to 63, for example, we could do so as follows:

```
x[1][3] = 63;
```

Observe how we use nested { } character pairs in the initialisation example above. In this case, the outer pair is associated with the entire array, while the three inner pairs are each associated with a row. In particular, observe how the elements in a row encapsulated by { } characters are comma-separated with no comma after the final element. Also, how the rows themselves are comma-separated with no comma after the final row. Once again, this can be confusing to beginners. And, once again, it all makes sense when you think about it.

Just for giggles and grins, we should probably note that it is also possible to initialise our array using a single pair of { } characters as follows:

```
int x[3][5] =
{
   10, 12, 99, 42, 45, 18, 15, 21,
   36, 54, 72, 91, 14, 62, 77
};
```

It's all Alan's fault

To be honest, I really hadn't intended to take such a diversion into the coding side of things. I know that our illustrious publisher, Matt Pulzer, is keen for us to perform as much nitty-gritty 'hands-on' experimentation as possible. However, everything we've discussed here will serve us well in the future.

Also, I think it's only fair to note that any blame for this excursion into the world of coding can be fairly and squarely laid at the feet of the aforementioned *PE* community member, Alan, who presented this poser in the first place.

Clickerty switch

OK, are you ready for some hands-on experiments? You are? Good. I am too. Two columns ago (*PE*, May 2023), I asked you to add a second switch to your breadboard (Fig.2). In our previous column (*PE*, June 2023), we added a passive piezoelectric buzzer and used it to present a more pronounced 'click' sound when we clicked our first switch. At the end of that column, I left you with two missions, the first being to add this click effect to our 'count up' code.

Just to make sure we are all tap-dancing in synchronisation, as it were, you can download a copy of our current breadboard layout showing both switches, our piezoelectric buzzer, our 7-segment display, and the connections to our Arduino Uno (file **CB-Jul23-01.pdf**). As usual, all of the files mentioned in this column are available from the July 2023 page of the *PE* website at: **https://bit.ly/pe-downloads**

After presenting our 'HELLO' message, our code currently commences by displaying 0. Pressing our first pushbutton causes the count to increment by 1. When the count reaches 9, the next press of the first pushbutton returns the count to 0. You can download the program containing the most recent incarnation of our code, including the `Click()` function that causes the piezoelectric buzzer to make our 'click' sound (file **CB-Jul23-02.txt**).

The main function of interest to us at the moment is our `GetNewDigit()` function (Listing 1). As we discussed in our previous column, a lot of this code is there to make sure our program doesn't inadvertently respond to switch bounce (ie, multiple events caused by the switch's electromechanical contacts making and breaking several times before settling in their new state).

Just to remind ourselves, the `PinSw` variable is associated with the Arduino's A0 pin, which is connected to the first switch. On Line 95 we read the value on this pin and store it in a local integer variable called `swVal`.

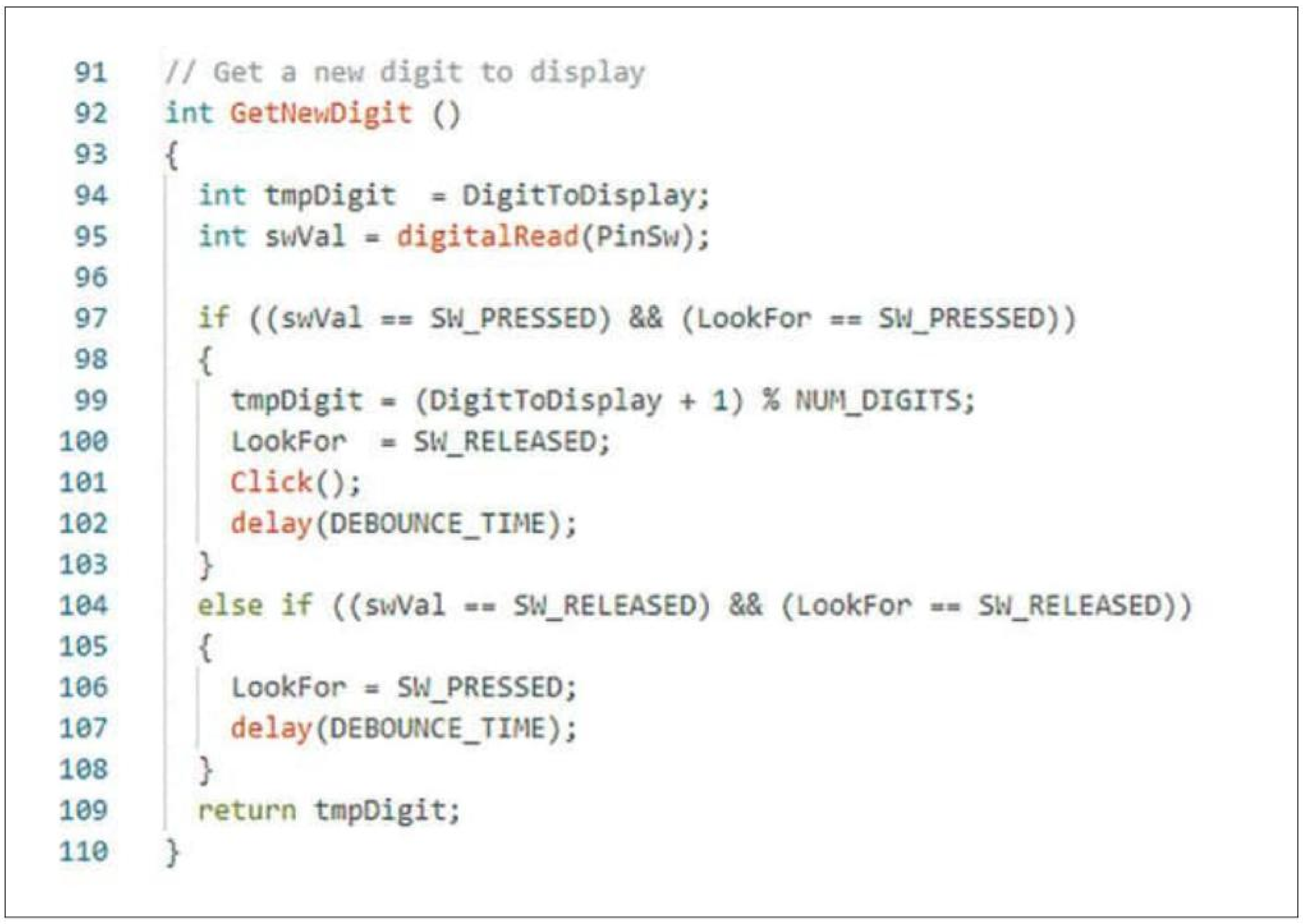
The portion of the code of interest to us here commences with the test on Line 97. If this test returns `true`, this means the switch has been pressed, in which case we execute the code in the associated compound statement (Lines 98 to 103). Look at us, saying things like 'compound statement' and actually having a clue what we're talking about – my mum is going to be so proud.

Observe the call to the `Click()` function on line 101. Also, observe Line 99, which is where the real magic takes place. The '`+ 1`' part is where we increment the count. The `% NUM_DIGITS` part is where we make sure that, when the count value reaches 10, it's returned to 0 again (we discussed this in ~~excruciating~~ exhilarating detail in *PE*, April 2023).

Crash and burn

This brings us to your second mission, which was to implement the code to get our second switch working. In this case, pressing the second pushbutton should cause the count to decrement by 1. Furthermore, if the count is at 0, then the next press of the second pushbutton should return the count to 9.

The first thing you should have done is replace the `PinSw` variable associated with the Arduino's A0 pin with two variables called something like `PinSw0` and `PinSw1`, which should be associated with pins A0 and A1, respectively. As a related item, in the `setup()` function you should have replaced the existing



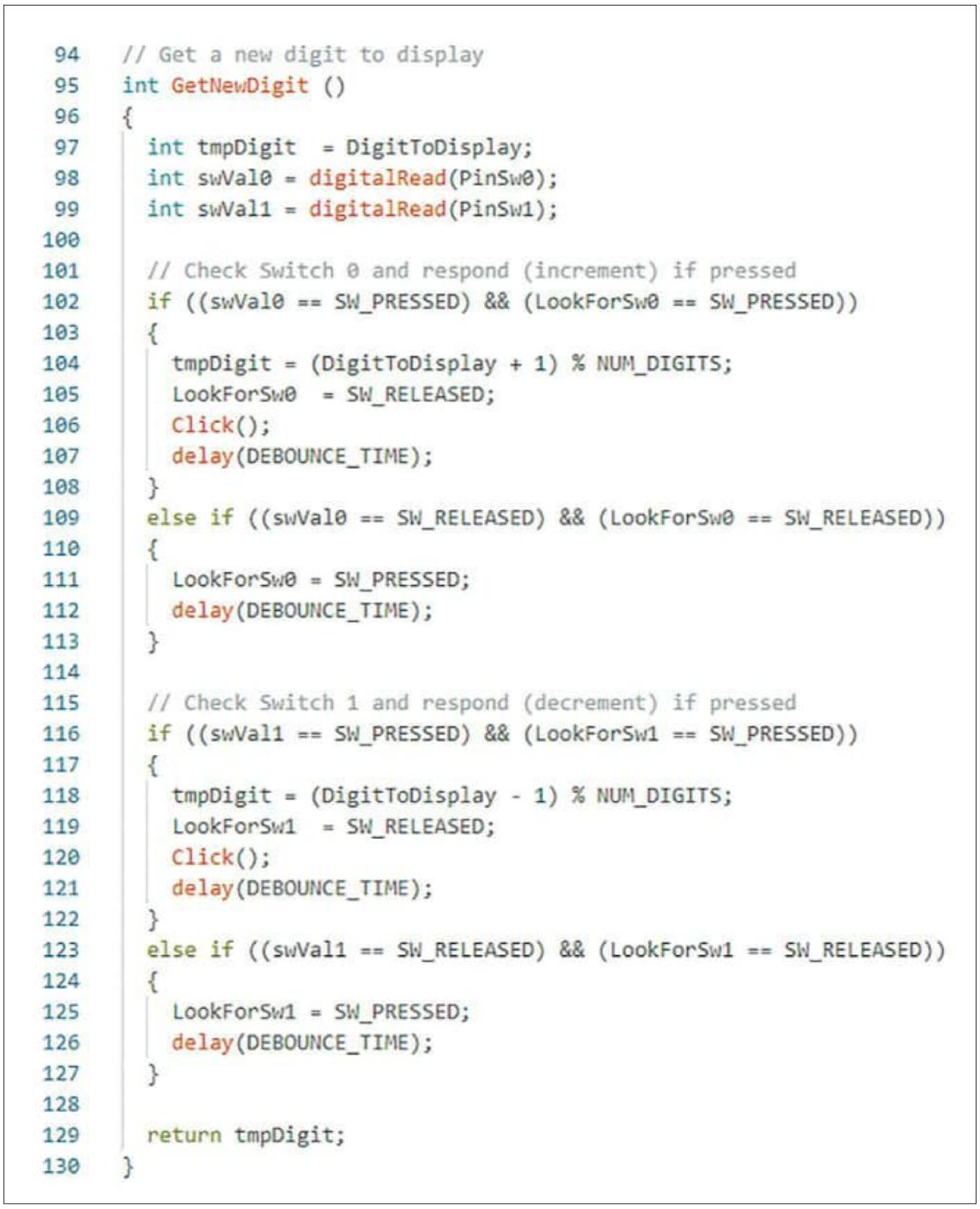
```
// Get a new digit to display
int GetNewDigit ()
{
  int tmpDigit  = DigitToDisplay;
  int swVal = digitalRead(PinSw);

  if ((swVal == SW_PRESSED) && (LookFor == SW_PRESSED))
  {
    tmpDigit = (DigitToDisplay + 1) % NUM_DIGITS;
    LookFor  = SW_RELEASED;
    Click();
    delay(DEBOUNCE_TIME);
  }
  else if ((swVal == SW_RELEASED) && (LookFor == SW_RELEASED))
  {
    LookFor = SW_PRESSED;
    delay(DEBOUNCE_TIME);
  }
  return tmpDigit;
}
```

Listing 1. Counting up.

`pinMode(PinSw, INPUT)` statement with two new statements reflecting the fact we now have two switches. Also, you should have replaced our global `LookFor` variable with two new variables called something like `LookForSw0` and `LookForSw1`.

Of course, your main task would have been to modify the `GetNewDigit()` function to accommodate both switches. For example, you may have captured some code looking something like that shown in Listing 2.



```
 94  // Get a new digit to display
 95  int GetNewDigit ()
 96  {
 97    int tmpDigit  = DigitToDisplay;
 98    int swVal0 = digitalRead(PinSw0);
 99    int swVal1 = digitalRead(PinSw1);
100
101    // Check Switch 0 and respond (increment) if pressed
102    if ((swVal0 == SW_PRESSED) && (LookForSw0 == SW_PRESSED))
103    {
104      tmpDigit = (DigitToDisplay + 1) % NUM_DIGITS;
105      LookForSw0  = SW_RELEASED;
106      Click();
107      delay(DEBOUNCE_TIME);
108    }
109    else if ((swVal0 == SW_RELEASED) && (LookForSw0 == SW_RELEASED))
110    {
111      LookForSw0 = SW_PRESSED;
112      delay(DEBOUNCE_TIME);
113    }
114
115    // Check Switch 1 and respond (decrement) if pressed
116    if ((swVal1 == SW_PRESSED) && (LookForSw1 == SW_PRESSED))
117    {
118      tmpDigit = (DigitToDisplay - 1) % NUM_DIGITS;
119      LookForSw1  = SW_RELEASED;
120      Click();
121      delay(DEBOUNCE_TIME);
122    }
123    else if ((swVal1 == SW_RELEASED) && (LookForSw1 == SW_RELEASED))
124    {
125      LookForSw1 = SW_PRESSED;
126      delay(DEBOUNCE_TIME);
127    }
128
129    return tmpDigit;
130  }
```

Listing 2. Counting up and down (first pass).

You can download this to peruse and ponder at your leisure (file **CB-Jul23-03.txt**). Actually, now that I come to think about it, it would have made more sense to have named our pins `PinSwUp` and `PinSwDn`, or at least add comments in the code saying which switch is supposed to do what, but I'm not going to go back and change things now (let's just treat this as a learning experience).

As usual, there are several things wrong with this code. Take a moment to see if you can spot any obvious areas of concern.

The first point is that we've totally divorced our handling of the two switches. We start by checking to see if our first switch has been pressed, in which case we respond accordingly, and then we move on to consider our second switch. What would happen if *both* switches were pressed at the same time? What *should* happen if both switches are pressed at the same time? Having both switches being activated or deactivated simultaneously may be unlikely (in some applications and implementations it may be impossible), but such a circumstance should at least be considered in the system's specification and – if necessary – accounted for in the code.

However, the real problem is located on Line 118, which is where we are attempting to decrement the count. As you can see, all we've done (I know it's me who did this, but I'm assuming you attempted something similar) is to take our increment count statement from Line 104 and replace the +1 with –1, which – as it turns out – is not a good idea.

The funny thing is that this will actually work some of the time, thereby making it a 'close, but no cigar' solution. Try running the program, wait for the initial '0' to appear on the 7-segment display, and then click the 'up' count button (our first switch) five times, thereby leaving '5' on the display. Now try clicking the 'down' count button (our second switch) five times, thereby returning the displayed value to '0'.

The problem arises when we click the 'down' count button one more time. We want this to take us to a displayed value of '9', but instead... Well, see what happens for yourself.

Don't be negative

What's gone wrong? Well, this is all to do with the way in which computers represent negative numbers. Do you remember when we introduced the concept of hexadecimal (*PE*, May 2023)? At that time, we noted that a 4-bit nybble can represent $2^4 = 16$ different patterns of 0s and 1s from 0000 to 1111, and that we can use these patterns to represent positive integers in the range 0 to 15. At that time, we failed to make mention of how we can also use these patterns to represent both positive *and* negative values. In fact, there are multiple methods available to us, two of the most common being known as 'sign-magnitude' and 'signed binary.'

Let's first remind ourselves that, in the case of an unsigned binary 4-bit nybble (Fig.3a), commencing with the rightmost bit, which is known as the least-significant bit (LSB), the column weights are $2^0 = 1$, $2^1 = 2$, $2^2 = 4$, and $2^3 = 8$. We determine the value of an unsigned binary number by multiplying each bit by its corresponding column weight and summing the results. For example, 0011 is $(0 \times 2^3) + (0 \times 2^2) + (1 \times 2^1) + (1 \times 2^0)$, which is the same as saying $(0 \times 8) + (0 \times 4) + (1 \times 2) + (1 \times 1) = 3$. Similarly, 1011 is $(1 \times 2^3) + (0 \times 2^2) + 1 \times 2^1) + (1 \times 2^0)$, which is the same as saying $(1 \times 8) + (0 \times 4) + (1 \times 2) + (1 \times 1) = 11$.

By comparison, in the case of sign-magnitude binary (Fig.3b), the leftmost bit, which is known as the most-significant bit (MSB) doesn't have a weight associated with it. Instead, it is used to represent the sign of the number, where 0 and 1 indicate positive and negative values, respectively. For this reason, not surprisingly, the MSB is also known as the 'sign bit.' Meanwhile, the three remaining bits continue to have the same weights as before. This means that, in the case of a nybble considered to represent sign-magnitude values, 0011 represents +3, while 1011 represents –(+3) = –3.

On the one hand, the sign-magnitude form is intuitive and balanced (we can use it to represent values from –7 to +7), and it is certainly employed for specific applications. On the other hand, we end up with two types of zero, –0 and +0, and we just know that this is going to be problematic at some stage in the game (what happens if we compare variables containing –0 and +0 values to see which is bigger, for example?).

And thus we come to signed binary (Fig.3c), which may appear to be a tad tricky at first, but which boasts an underlying elegance and sophistication (I'm a big fan). In this case, the MSB represents a negative quantity, which is $-2^3 = -8$ in the case of a 4-bit nybble. Once again, the three remaining bits continue to have the same positive weights as before. On the one hand this results in an unbalanced range of –8 to +7. On the other hand, we now have only a single zero, which has to make our lives easier in the long run.

Let's think this through. In this case, 0011 is $(0 \times -2^3) + (0 \times 2^2) + (1 \times 2^1) + (1 \times 2^0)$, which is the same as saying $(0 \times -8) + (0 \times 4) + (1 \times 2) + (1 \times 1) = 3$. However, 1011 is $(1 \times -2^3) + (0 \times 2^2) + 1 \times 2^1) + (1 \times 2^0)$, which is the same as saying $(1 \times -8) + (0 \times 4) + (1 \times 2) + (1 \times 1)$, which boils down to $-8 + 3 = -5$.

If this is the first time you are seeing this, you may be forgiven for thinking 'What?' and 'What??' and even 'What???' At first glance, signed binary might appear to be an exceedingly convoluted solution to a relatively trivial problem. All I can ask is that you stand firm and keep the faith because there is much more to the signed binary format than meets the eye, not least that it dramatically eases the way in which we assemble the logic functions that allow a digital computer to perform its magic.

Components from Part 1

LEDs (assorted colours)	**https://amzn.to/3E7VAQE**
Resistors (assorted values)	**https://amzn.to/3O4RvBt**
Solderless breadboard	**https://amzn.to/3O2L3e8**
Multicore jumper wires (male-male)	**https://amzn.to/3O4hnxk**

Components from Part 2

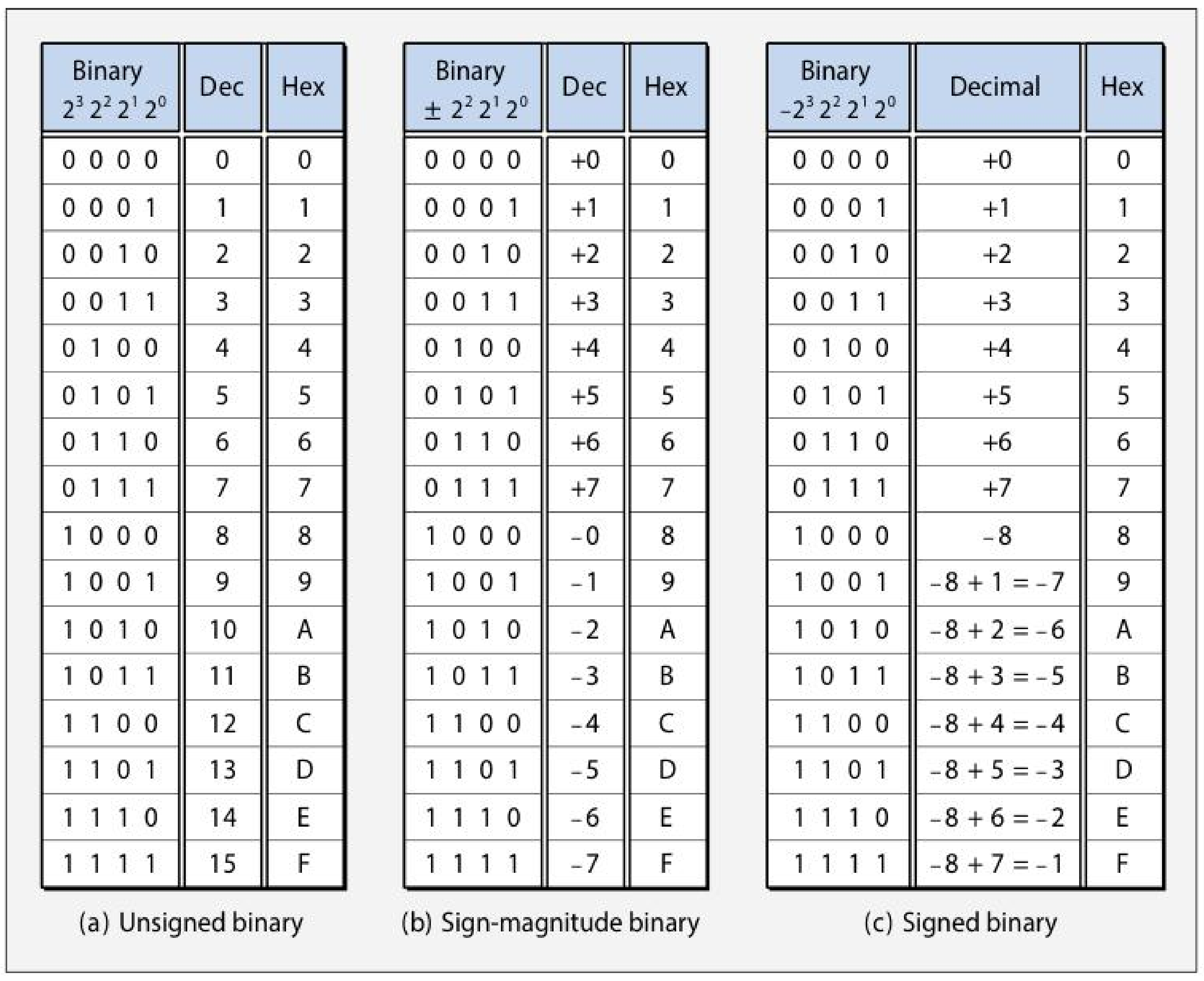
7-segment display(s)	**https://amzn.to/3Afm8yu**

Components from Part 5

Momentary pushbutton switches	**https://amzn.to/3Tk7Q87**

Components from Part 6

Passive piezoelectric buzzer	**https://amzn.to/3KmxjcX**



Binary $2^3\ 2^2\ 2^1\ 2^0$	Dec	Hex
0 0 0 0	0	0
0 0 0 1	1	1
0 0 1 0	2	2
0 0 1 1	3	3
0 1 0 0	4	4
0 1 0 1	5	5
0 1 1 0	6	6
0 1 1 1	7	7
1 0 0 0	8	8
1 0 0 1	9	9
1 0 1 0	10	A
1 0 1 1	11	B
1 1 0 0	12	C
1 1 0 1	13	D
1 1 1 0	14	E
1 1 1 1	15	F

(a) Unsigned binary

Binary $\pm\ 2^2\ 2^1\ 2^0$	Dec	Hex
0 0 0 0	+0	0
0 0 0 1	+1	1
0 0 1 0	+2	2
0 0 1 1	+3	3
0 1 0 0	+4	4
0 1 0 1	+5	5
0 1 1 0	+6	6
0 1 1 1	+7	7
1 0 0 0	–0	8
1 0 0 1	–1	9
1 0 1 0	–2	A
1 0 1 1	–3	B
1 1 0 0	–4	C
1 1 0 1	–5	D
1 1 1 0	–6	E
1 1 1 1	–7	F

(b) Sign-magnitude binary

Binary $-2^3\ 2^2\ 2^1\ 2^0$	Decimal	Hex
0 0 0 0	+0	0
0 0 0 1	+1	1
0 0 1 0	+2	2
0 0 1 1	+3	3
0 1 0 0	+4	4
0 1 0 1	+5	5
0 1 1 0	+6	6
0 1 1 1	+7	7
1 0 0 0	–8	8
1 0 0 1	–8 + 1 = –7	9
1 0 1 0	–8 + 2 = –6	A
1 0 1 1	–8 + 3 = –5	B
1 1 0 0	–8 + 4 = –4	C
1 1 0 1	–8 + 5 = –3	D
1 1 1 0	–8 + 6 = –2	E
1 1 1 1	–8 + 7 = –1	F

(c) Signed binary

Fig.3. Two ways of representing negative numbers in binary.

More than a nybble

Did you observe that the hexadecimal values remain unchanged, irrespective of whether we are considering our nybble to represent unsigned, sign-magnitude, or signed binary values (Fig.3)? This is because hexadecimal values are no more than a one-to-one mapping to their binary counterparts.

But what about values larger than a 4-bit nybble (Fig.4a)? Well, in the case of an 8-bit byte (Fig.4b), the MSB would represent $-2^7 = -128$, while all the other bits continue to have positive weights. In turn, this means an 8-bit signed binary number can be used to represent values in the range –128 to +127. Similarly, in the case of a 16-bit word (Fig.4c), the MSB would represent $-2^{15} = -32{,}768$, which means a 16-bit signed binary number can be used to represent values in the range –32,768 to +32,767.

The big point here is that, in the case of the signed binary format, all 1s values always represent –1. By comparison, if we were considering these fields to represent unsigned binary values, then their MSBs would represent $+2^3$, $+2^7$, and $+2^{15}$, and their all 1s values would represent 15, 255, and 65,535 for our nybble, byte, and word examples, respectively.

Ah, that explains it!

This is probably a good time to mention that, in the case of the Arduino Uno microcontroller which we are using for our experiments, the `int` (integer) data type is a 16-bit field. Also, we can declare integers as being signed (ie, signed binary) or unsigned, for example:

```
signed   int a;
unsigned int b;
```

If we simply declare a variable as being of type `int`, then the compiler will treat it as being signed by default. The really

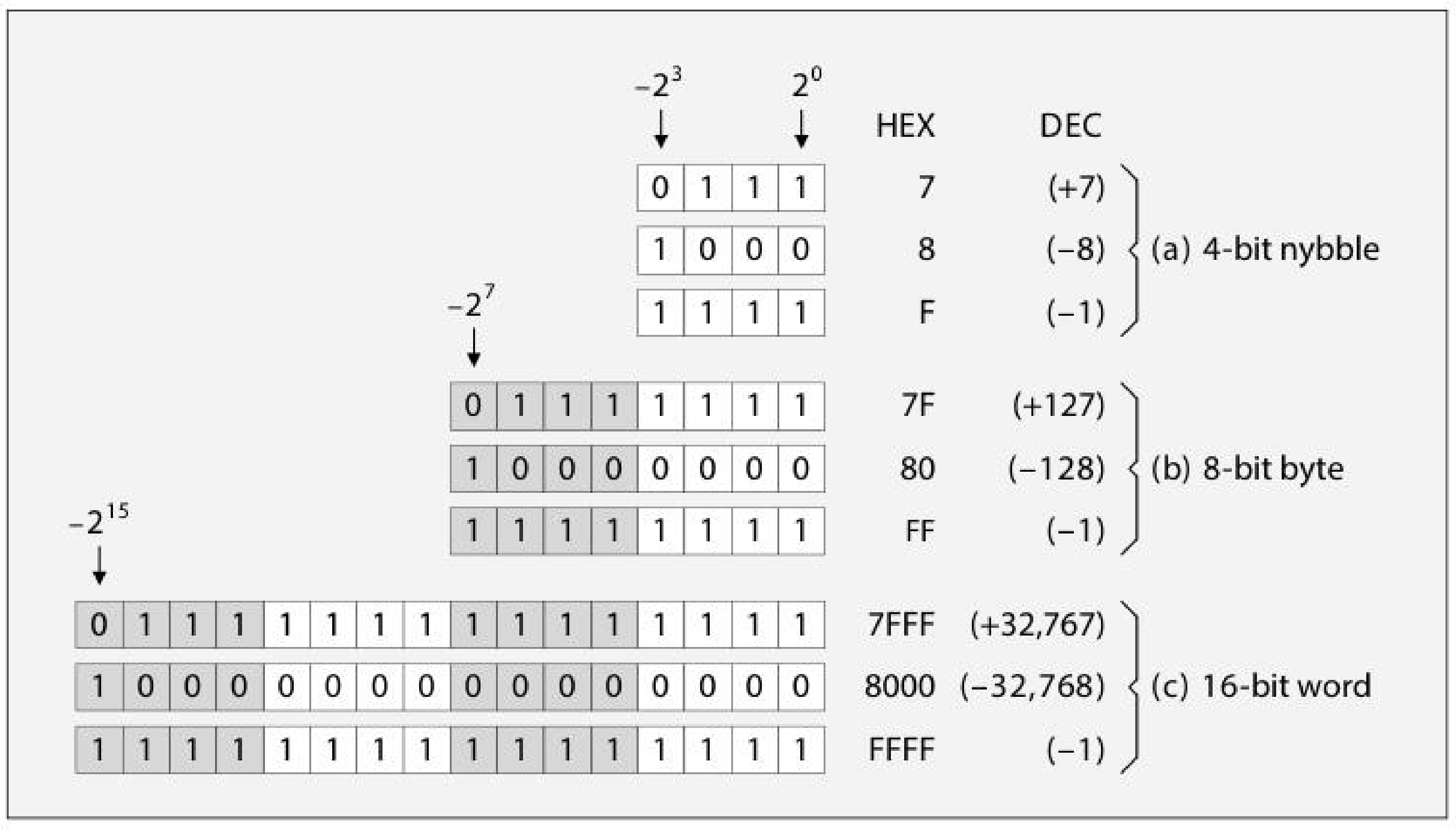


Fig.4. Signed binary nybbles, bytes, and words.

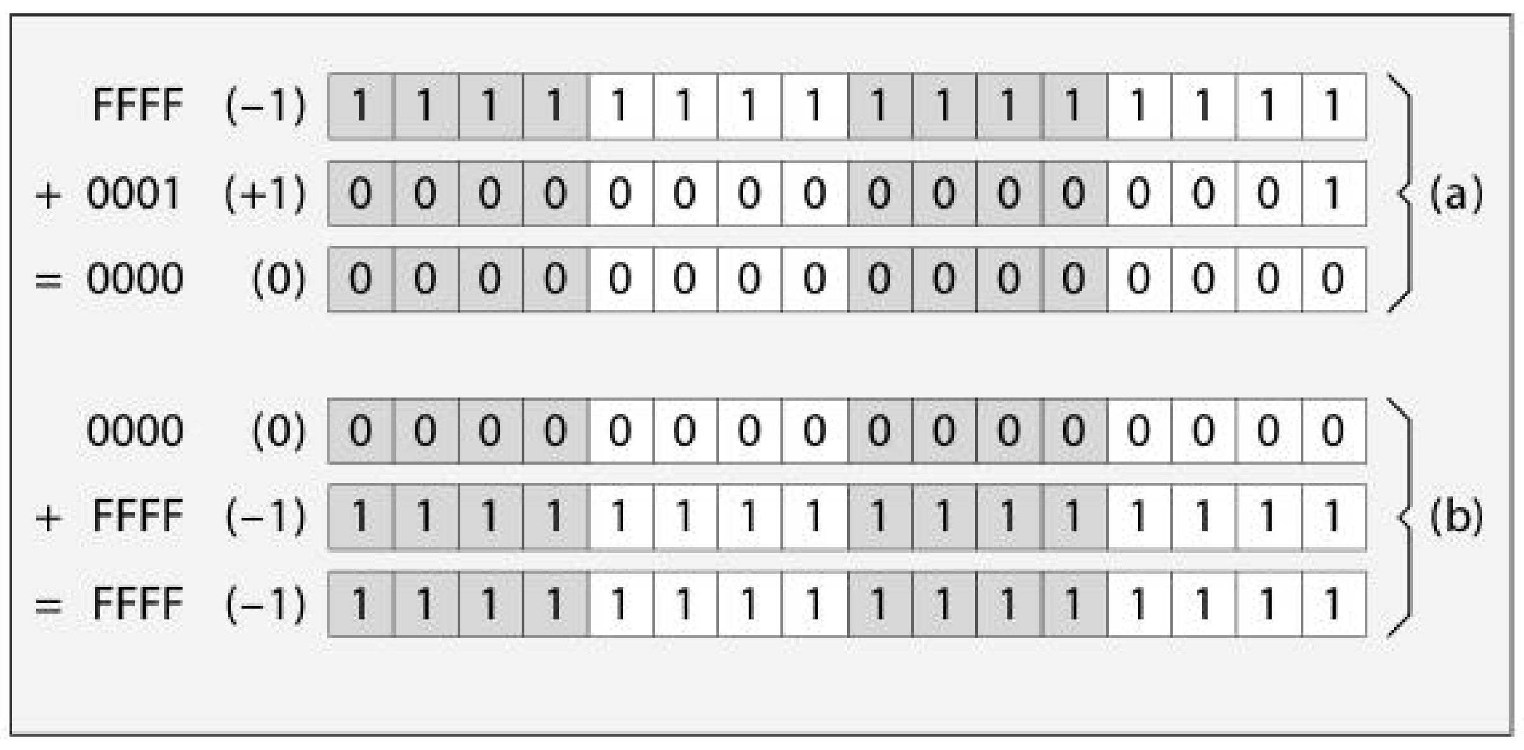


Fig.5. Operations of interest with signed binary.

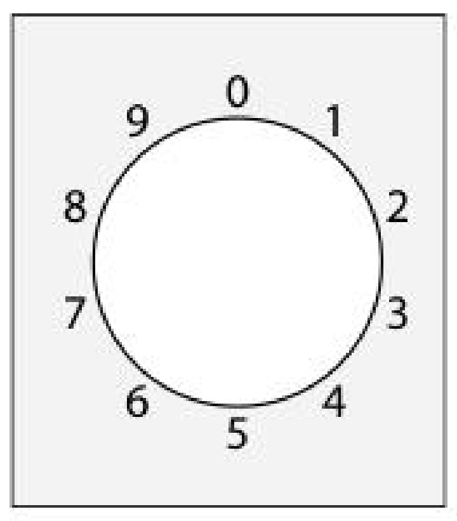


Fig.6. Circular reasoning.

interesting thing is that signed and unsigned integers are stored and manipulated inside the computer in the same way – the only thing that's different is the values we consider them to represent.

We're almost there. The penultimate thing to consider is what is going to happen if an integer variable already contains all 1s and we add 1 to it. In this case, whatever 16-bit hardware (register, memory location...) is being used to store this value will overflow and return to containing all 0s, irrespective of whether we are considering it to represent signed or unsigned values. Bearing this in mind, if we are considering our variable to represent signed binary values, which means an all 1s value represents –1, it should come as no surprise that –1 + 1 = 0 (Fig.5a).

Contrariwise, what happens if our variable contains all 0s and we subtract one from it. In this case we have to remember that 0 – 1 is the same as saying 0 + (– 1). And, once again, since we know that all 1s represents –1, it should come as no surprise that 0 + (–1) = –1 (Fig.5b).

All of this goes to explain why our counting down code went pear-shaped. The easiest way to wrap our brains around all of this is to employ a little circular reasoning. Let's envisage our ten digits (0 to 9) as being arranged in a circle (Fig.6).

Let's start by reminding ourselves what happens when we count up using the statement on Line 104 of Listing 2 (remember that the modulo `%` operator returns the remainder from an integer division and that we've defined `NUM_DIGITS` to have a value of 10):

```
tmpDigit = (DigitToDisplay + 1) % NUM_DIGITS;
```

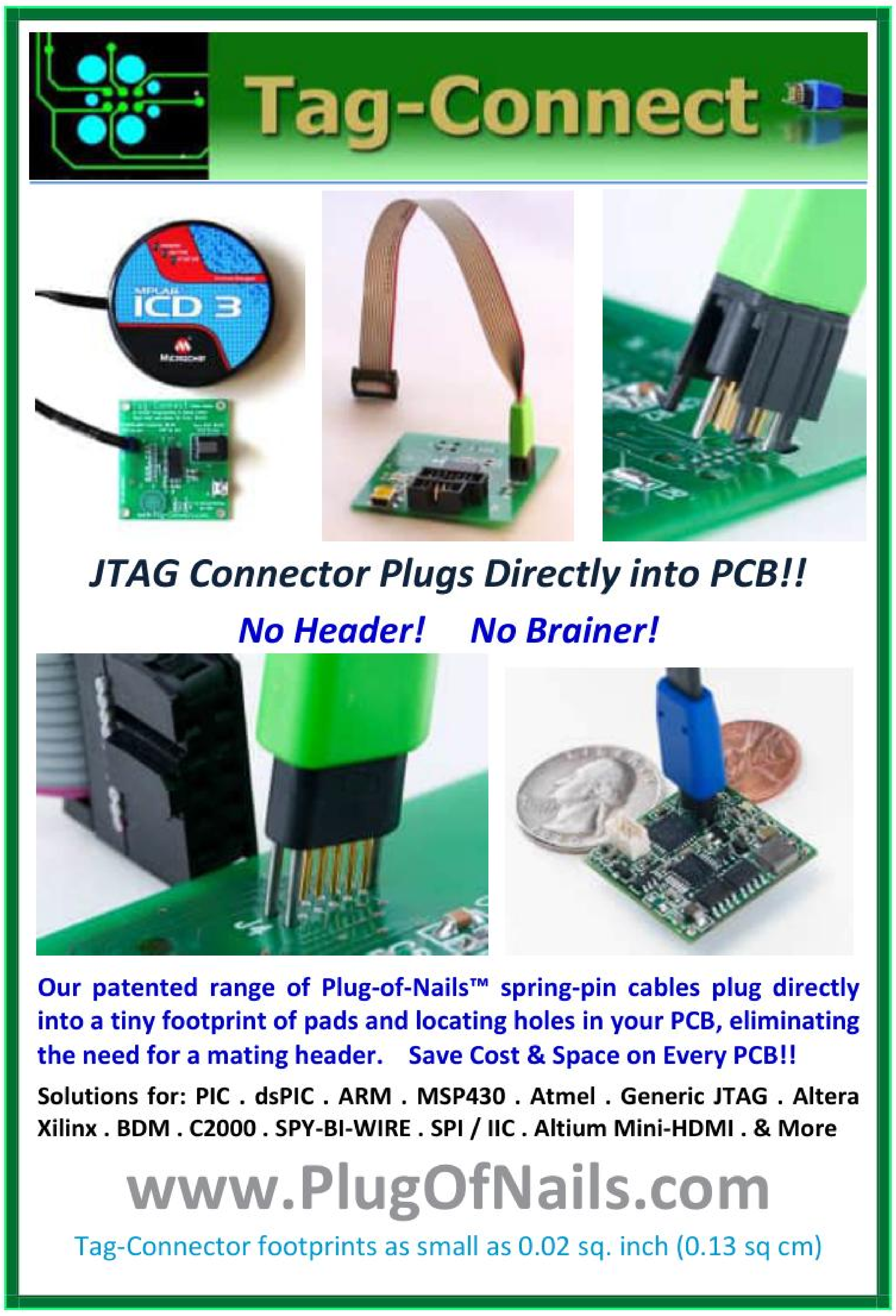
If `DigitToDisplay` is 0 to 8, adding 1 will take us in a clockwise direction, returning 1 to 9, respectively, and this is the remainder value that will be returned by the `% 10` (ie, `% NUM_DIGITS`) operation. By comparison, when `DigitToDisplay` is 9, adding 1 will give us 10, and `10 % 10` will return 0, which is what we want.

Now, let's consider what happens when we count down using the statement on Line 118 of Listing 2:

```
tmpDigit = (DigitToDisplay - 1) % NUM_DIGITS;
```

In this case, if `DigitToDisplay` is 9 to 1, subtracting 1 will take us in a anticlockwise direction, returning 8 to 0, respectively, and this is the remainder value that will be returned by the `% 10` operation. By comparison, when `DigitToDisplay` is 0, subtracting 1 will give us -1, but this is represented by all 1s in signed binary, and your guess is as good as mine as to what the `% 10` operation will do with this (suffice it to say that it won't return a value of 9).

Happily, there is a solution. Suppose we start by adding 10 to whatever value is currently stored in `DigitToDisplay`, which will take us a full turn clockwise round the dial in Fig.6. This means our updated version Line 118 in Listing 2 will be as follows (file **CB-Jul23-04.txt**):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```
tmpDigit = (DigitToDisplay + NUM_DIGITS - 1)
             % NUM_DIGITS;
```

In this case, if `DigitToDisplay` is 9 to 1, adding 10 will give 19 to 11, respectively; subtracting 1 will give 18 to 10, respectively; and performing `% 10` will result in 8 to 0 as before. Furthermore, when `DigitToDisplay` is 0, adding 10 will give 10, subtracting 1 will give 9, and performing `% 10` will result in a remainder value of 9, which is what we want.

As usual, I feel a little 'Tra-la' is in order.

Homework

I'm sorry we've spent so much of our time in this column contemplating things that make our heads ache, but everything we've learned here is going to come in very, very useful in the future.

What I'd like you to do now is to think of at least three different applications we can implement with our setup so far, which is two buttons, a piezoelectric buzzer, and our 7-segment display. I'll even start the ball rolling by suggesting one possible application, which would be to implement a form of electronic dice. Every time we press a button, our program could display a random number from 1 to 6.

How about we jazz this up a little by generating a series of random numbers before settling on the final value? What about making the random numbers change quickly at first and then gradually slow down before eventually settling on a final value? Might we want to arrange things such that each number is accompanied by a click on our buzzer and a merry little jingle plays when the final value is displayed?

This counts as just one application. I'm sure you can come up with more (feel free to email me to let me know what you're thinking, and I'll respond by sharing my own ideas). Until next time, have a good one!

Cool bean Max Maxfield (Hawaiian shirt, on the right) is emperor of all he surveys at **CliveMaxfield.com** – the go-to site for the latest and greatest in technological geekdom.

Comments or questions? Email Max at: **max@CliveMaxfield.com**

AUDIO OUT

By Jake Rothman

Potentiometer scratching – Part 2

Last month, we introduced the problem of 'pot scratching'. An annoying fault that causes audio output to go 'scratch, crackle and bang'. This month, we'll examine how to tackle the problem and offer some pot recommendations.

Measurements and testing

The *Radio Designer's Handbook* states that a hot-moulded carbon track pot will typically give 1-2mV/V applied across the track when rotated at a speed of one rotation per second. I made some measurements and found that a current of just 1.5µA through the wiper could just make audible noise on a 25kΩ linear Plessey MCH 0.5W moulded-carbon pot listened to at line-level. This is a problem, because most electrolytic capacitors have a leakage of around 3 to10µA. On the other side of the pot feeding the clockwise end of the track, 5µA could be tolerated. This experimental set-up is shown in Fig.11.

Oxidation

Oxidation of the metal components of potentiometers, such as the wiper and its central pick up contact (slip ring) is a particular problem in damp (eg, British) climates. The modular Allen-Bradley ModPot 70 series is very successful in the US market, but fails from wiper slip-ring assembly oxidation in British humidity. (The wiper itself is immune, being made from carbon as shown in Fig.12 – it's the rest of the device that oxidises).

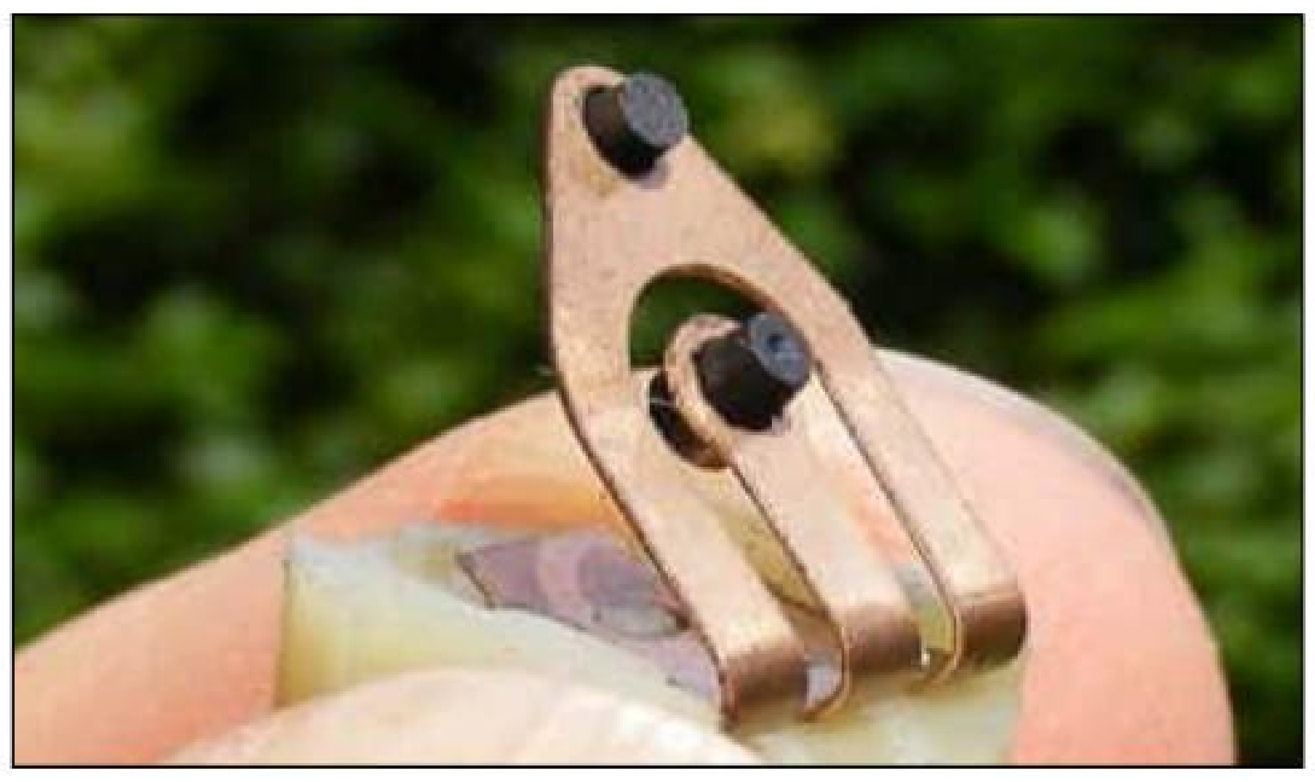


Fig.12. Most moulded-track pots have carbon brush wipers, such as this Plessey MH1. This can reduce rotational noise compared to a metal wiper.

The otherwise excellent Wayne Kerr AMS1 audio measurement system (Fig.13) suffers from scratchy devices made by ModPots. This is not helped by the lack of DC blocking capacitors in the circuit. This was further exacerbated by the use of NE5534 op amps, which have high input bias currents and output offsets. Furthermore, the system's thermistor-stabilised phase-shift oscillator configuration (Fig.14) is one of the most sensitive circuits to pot noise, which causes the amplitude to bounce all over the place as the frequency is altered. Another problem is that the anti-log dual-gang 10kΩ pot used has an unusual 20% taper. Changing it to another manufacturer makes the scale markings wrong. Last, but not least, the other pot configurations of tracks and switches are also unique and the only solution is total disassembly and servicing. Luckily ModPots are held together with screws as shown in Fig.15, so access to the innards is possible.

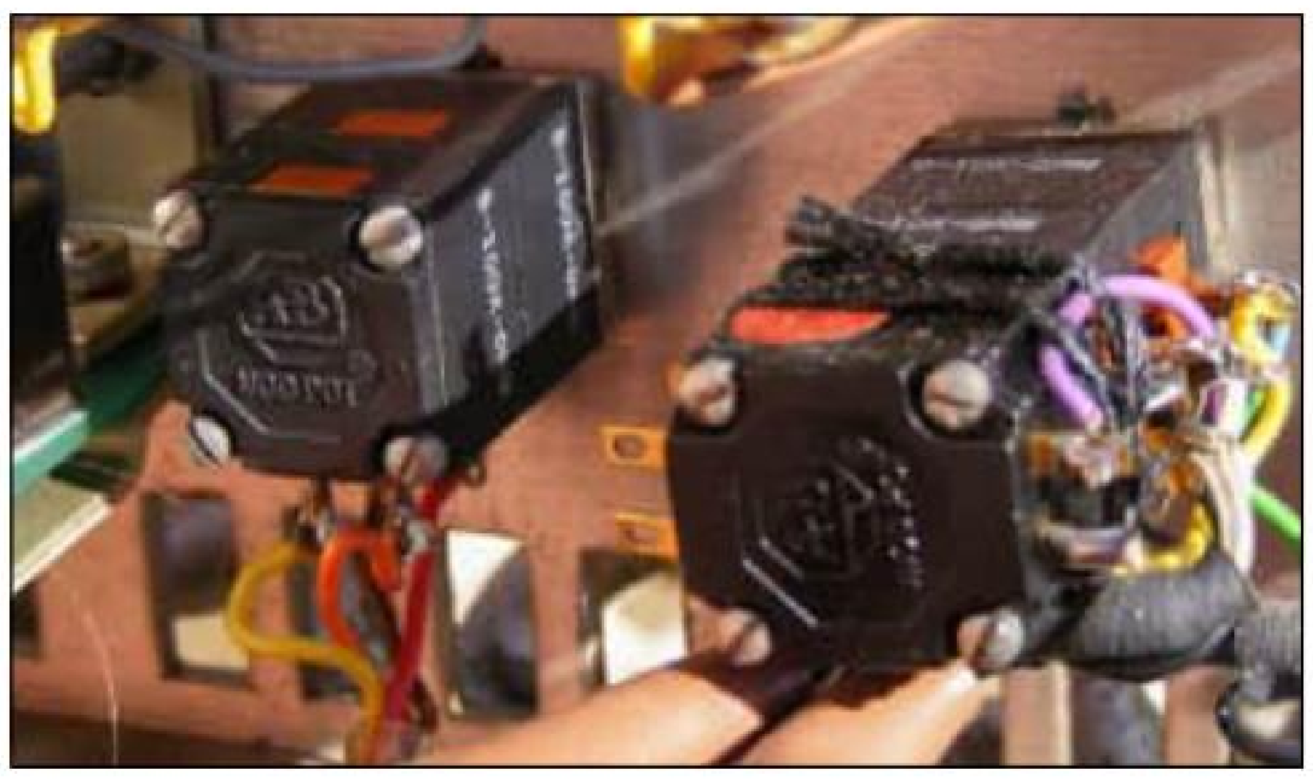


Fig.15. Luckily for service engineers, the AMS1's custom ModPots are screwed together, facilitating servicing.

When I was responsible for maintaining Solid State Logic 4000G series mixing desks I often stripped down and fixed their unique multi-section Clarostat 388 series pots. At £30 a device in the 1980s, it was worth it.

Old Omeg pots which used Radiohm parts had a peculiar slip-ring problem because the wiper is brass and the centre pin contact is steel (see Fig.16). The dissimilar metals reacted badly in damp and sometimes even acted like a rectifier. Around 1993 the centre ring was replaced

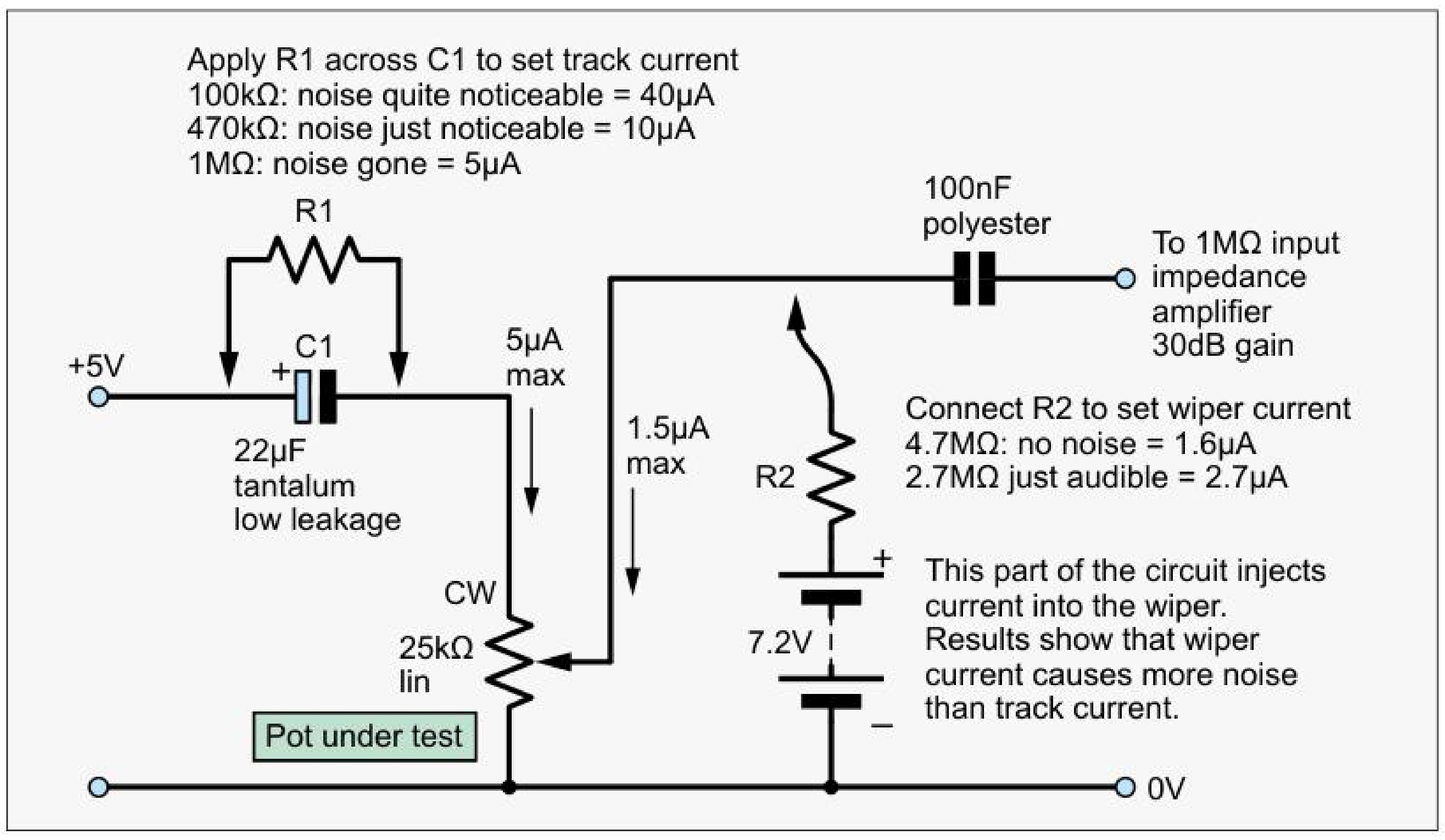

Apply R1 across C1 to set track current
100kΩ: noise quite noticeable = 40µA
470kΩ: noise just noticeable = 10µA
1MΩ: noise gone = 5µA
R1
C1
+5V
22µF
tantalum
low leakage
5µA
max
1.5µA
max
CW
25kΩ
lin
Pot under test
100nF
polyester
To 1MΩ input
impedance
amplifier
30dB gain
Connect R2 to set wiper current
4.7MΩ: no noise = 1.6µA
2.7MΩ just audible = 2.7µA
R2
7.2V
This part of the circuit injects
current into the wiper.
Results show that wiper
current causes more noise
than track current.
0V

Fig.11. The maximum currents allowed through pots before rotational noise becomes apparent are very small, as shown in this experiment.

PHONES
VOLUME
SCOPE
DISTORTION
OUTPUT
O/P LEVEL
FREQUENCY (Hz)

Fig.13. This old Wayne Kerr AMS1 audio analyser was cursed with scratchy pots and clicky switches.

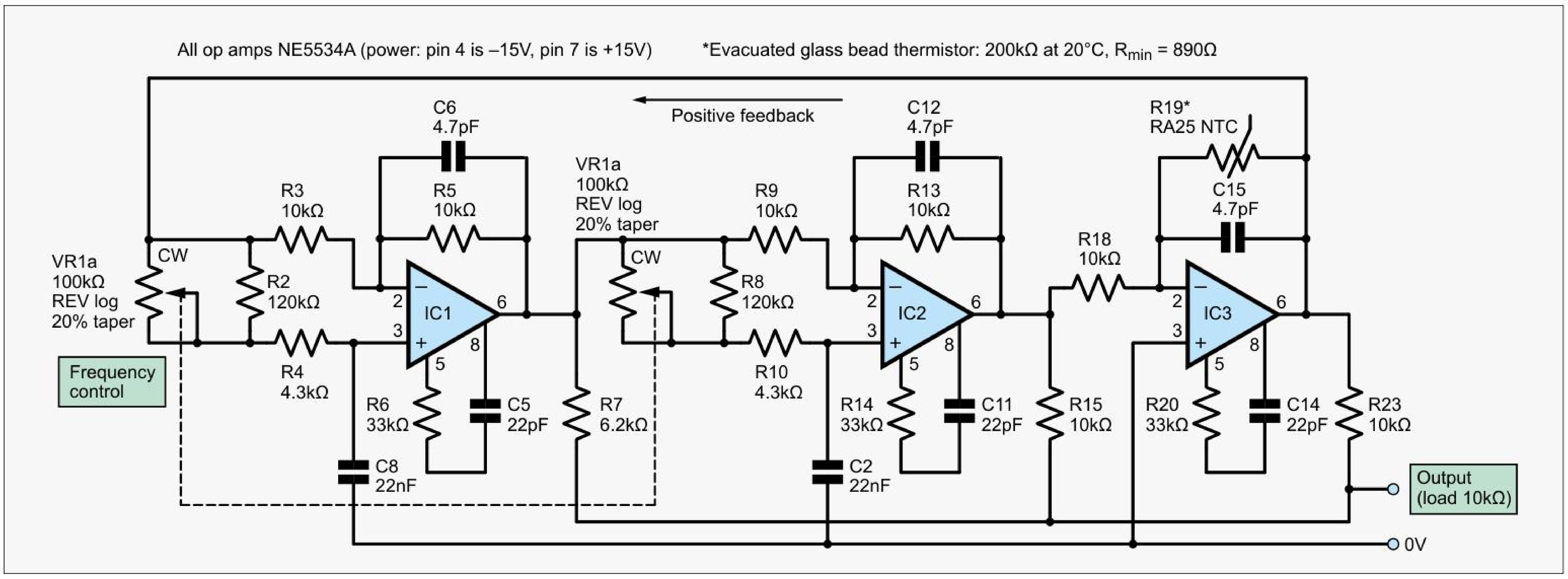

All op amps NE5534A (power: pin 4 is –15V, pin 7 is +15V)
*Evacuated glass bead thermistor: 200kΩ at 20°C, R$_{min}$ = 890Ω
Positive feedback
VR1a 100kΩ REV log 20% taper
CW
Frequency control
R3 10kΩ
R2 120kΩ
R4 4.3kΩ
C6 4.7pF
R5 10kΩ
IC1
R6 33kΩ
C5 22pF
C8 22nF
VR1a 100kΩ REV log 20% taper
CW
R7 6.2kΩ
R9 10kΩ
R8 120kΩ
R10 4.3kΩ
C12 4.7pF
R13 10kΩ
IC2
R14 33kΩ
C11 22pF
C2 22nF
R15 10kΩ
R18 10kΩ
R19* RA25 NTC
C15 4.7pF
IC3
R20 33kΩ
C14 22pF
R23 10kΩ
Output (load 10kΩ)
0V

Fig.14. The AMS1's low-distortion oscillator circuit suffered from pot noise. It can be fixed by using 10µF polyester DC-blocking capacitors in series with the pots.

with a printed carbon polymer track (Fig.17) and the problem was solved. (This caused me constant problems with Colorsound guitar pedals and I dislike sourcing parts from overseas for a product with such a unique British history. I'll be going back to UK sourcing from Omeg.)

Servicing pots

A common cause of pot noise is dirt on the track causing intermittent contact. There is also the build-up of carbon from track wear. Pots tend to be open or sealed. Open types are more likely to become noisy but they can easily be cleaned by squirting cleaner through the gap above the tags. Sealed ones can be quite a challenge to open non-destructively. A switch contact cleaner lubricant aerosol, such as Super Servisol 10 (now marketed as Kontakt super 10) is the standard cleaner. This is available from most suppliers such as Cricklewood Electronics. The very expensive Contralube 770 connector protector from Newgate Simms Ltd (see Fig.18) is an excellent synthetic gel pot track lubricant. I sourced my 75g tube from Rapid, but sadly it now seems to be discontinued. It's still possible to get 8g packs for £8.00 from specialist automotive suppliers, and there is an alternative compound called Tribogel.

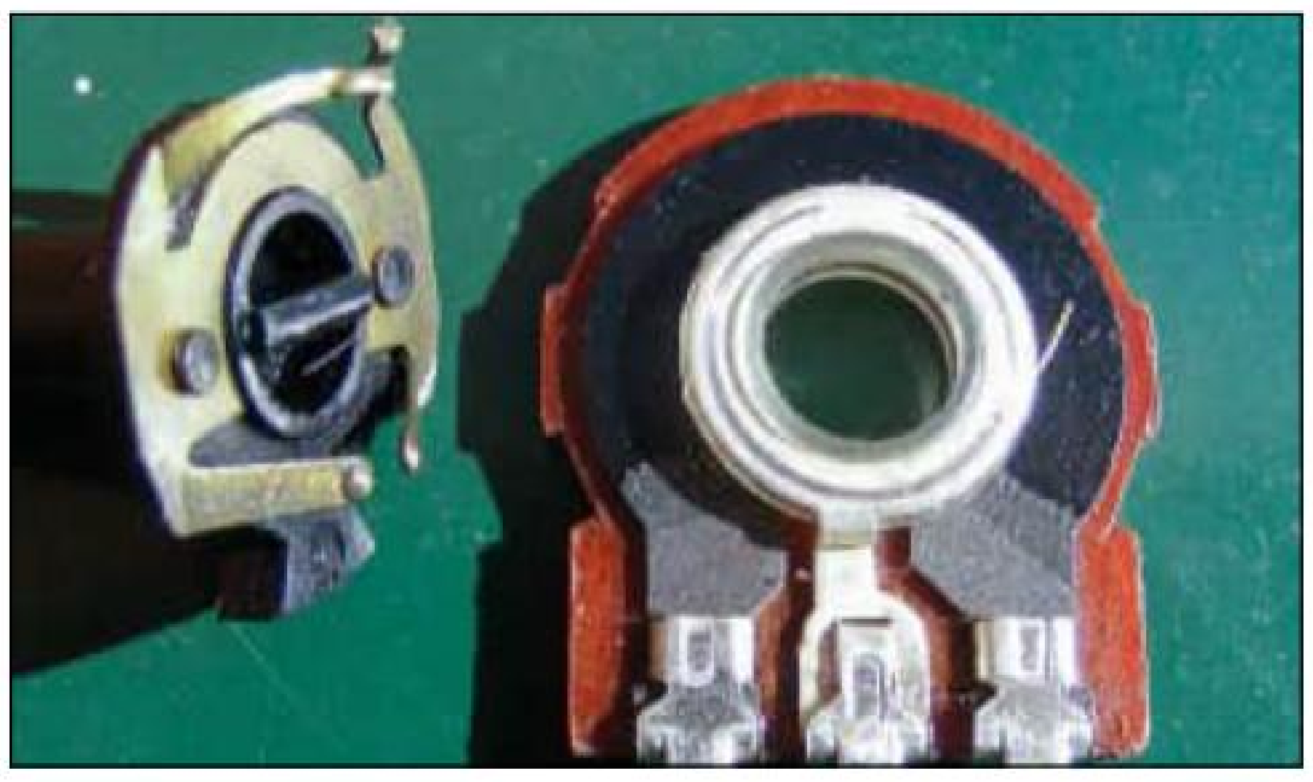


Fig.16. Old Radiohm/Omeg P20 (pre-1993) pots had particularly crackle-prone centre terminal slip-ring assemblies due to dissimilar metals being used.

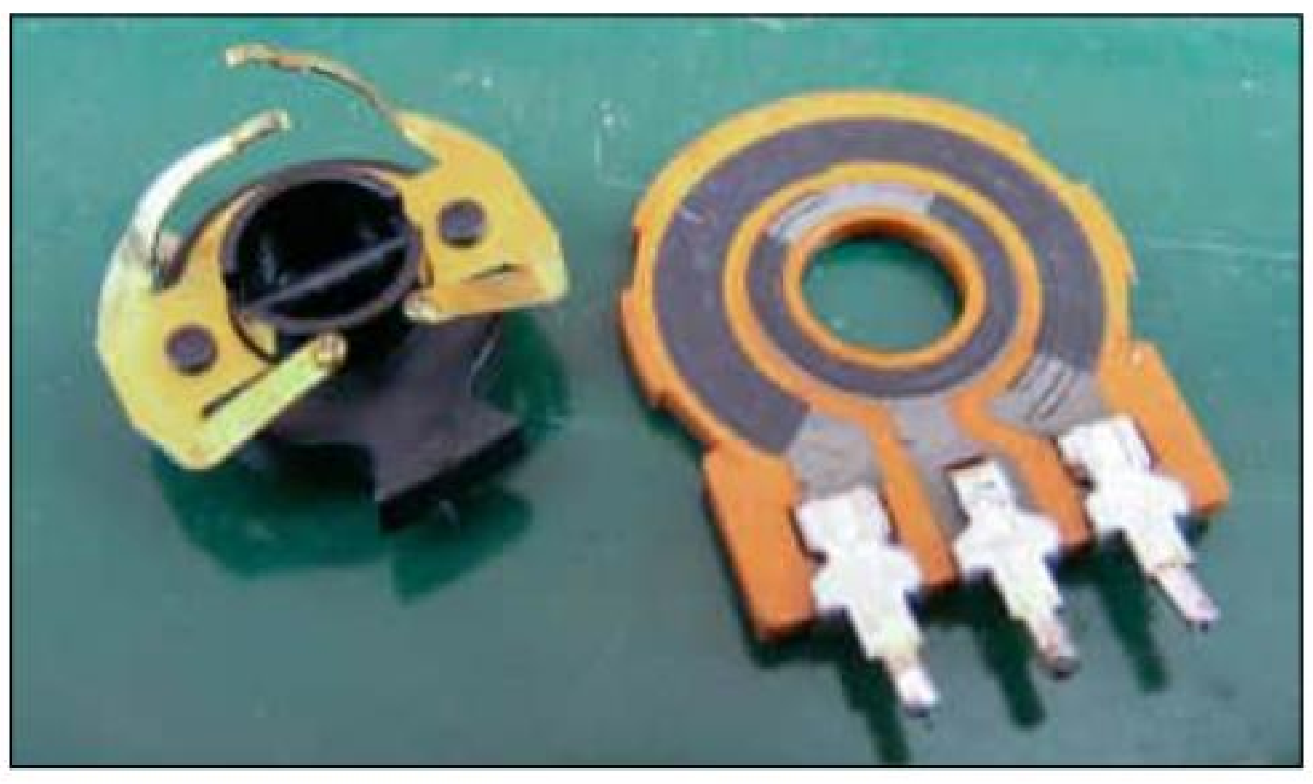


Fig.17. Later Omeg P20 pots had a carbon polymer track and centre slip-ring, which eliminated the above failure mode and were much quieter.

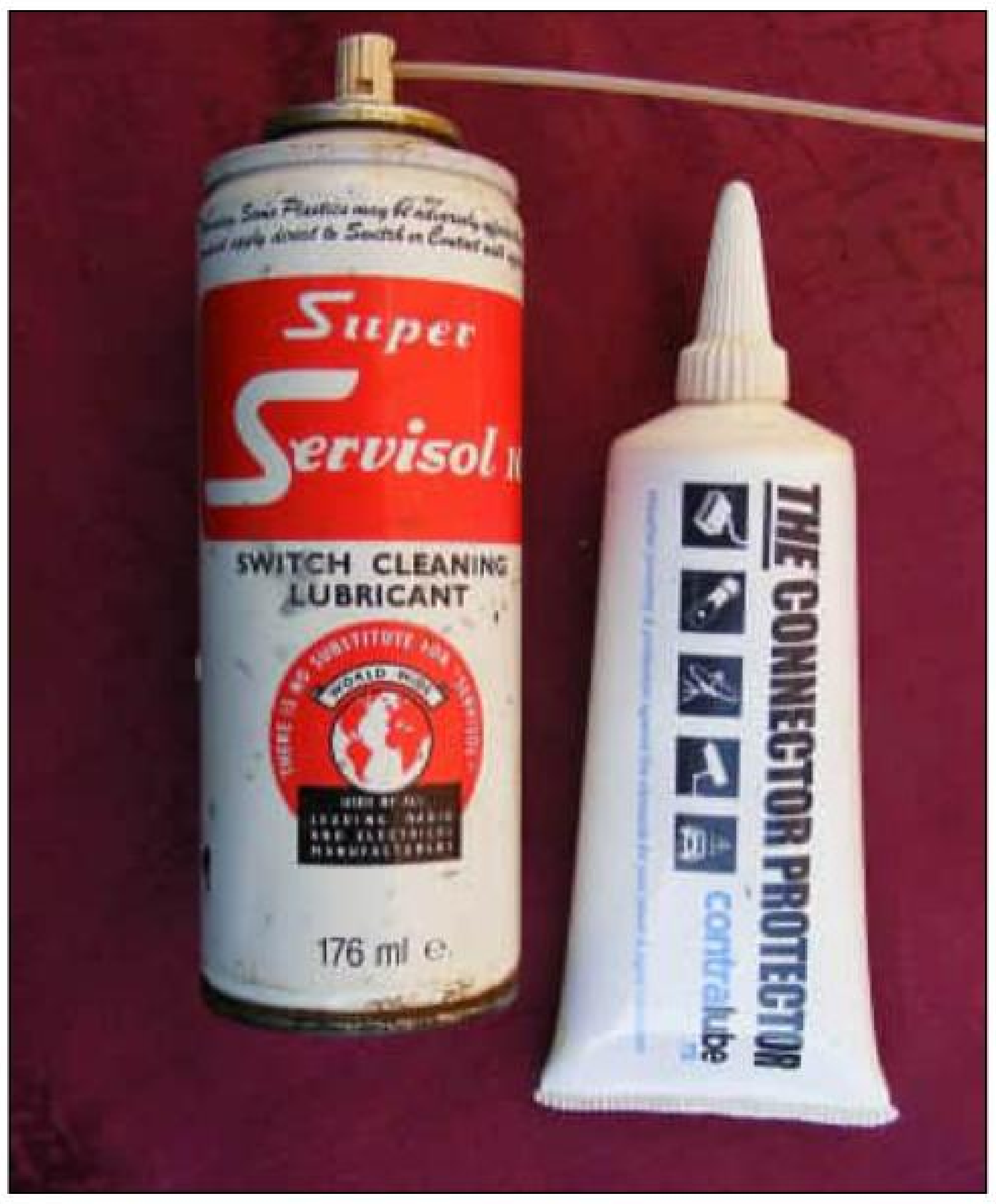

Super Servisol 10
SWITCH CLEANING LUBRICANT
176 ml ℮
THE CONNECTOR PROTECTOR
contralube

Fig.18. Super Servisol and Contralube, old mainstays of many a studio maintenance engineer's tool kit.

Don't use the ubiquitous WD40 spray, since its solvents soon evaporate, leaving an oil film not dissimilar to 3-in-1 mineral oil. This then turns to a sticky resin over time. It is meant for freeing seized bolts and locks, not cleaning pots and switches, although it's useful for loosening stiff pot shaft bushings. Newgate Simms also offer mechanical damping grease for pot bushes. If you want that 'velvety feel' to pot rotation then a low-viscosity type is recommended. Kilopoise grease can be too stiff, but it's what Omeg used to use.

There are reactive cleaners that claim to dissolve metallic oxides and sulphides. DeoxIT F5 by Caig Chemicals in the US is possibly the best known compound. They also do a special Fader Lube for conductive plastic pots. These products are distributed by Hot Rox UK in Nottingham. I use the tube versions of the liquid shown in Fig.19 and their wipes (Fig.20) which require the pot to be disassembled, unlike the aerosol options.

Do remember that most aerosols are around 80% propellant and are therefore an expensive way of dispensing chemicals. They also pollute the workshop air. One oxide-removing aerosol I do recommend however is Kontakt 60 (Rapid 01-0744) since it is relatively cheap. Remember to re-lubricate afterwards to prevent rapid re-oxidation.

These chemicals often need a bit of extra help when dealing with oxidised wiper assemblies and a little abrasion is needed.

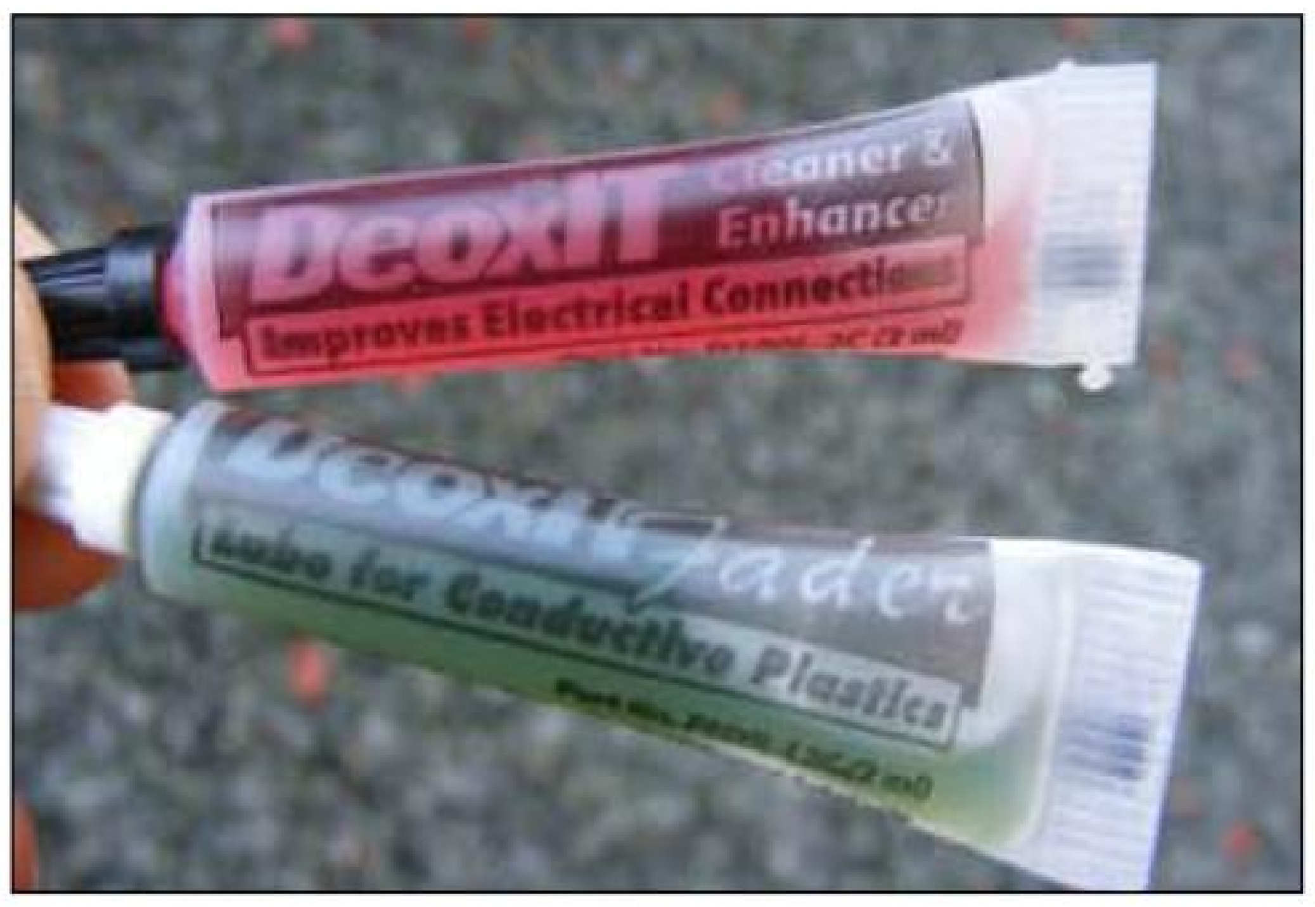

DeoxIT Cleaner & Enhancer
Improves Electrical Connections
DeoxIT Fader
Lube for Conductive Plastics

Fig.19. Specialist products such as this DeoxIT fader lubricant can be very useful if you do a lot of pot maintenance.

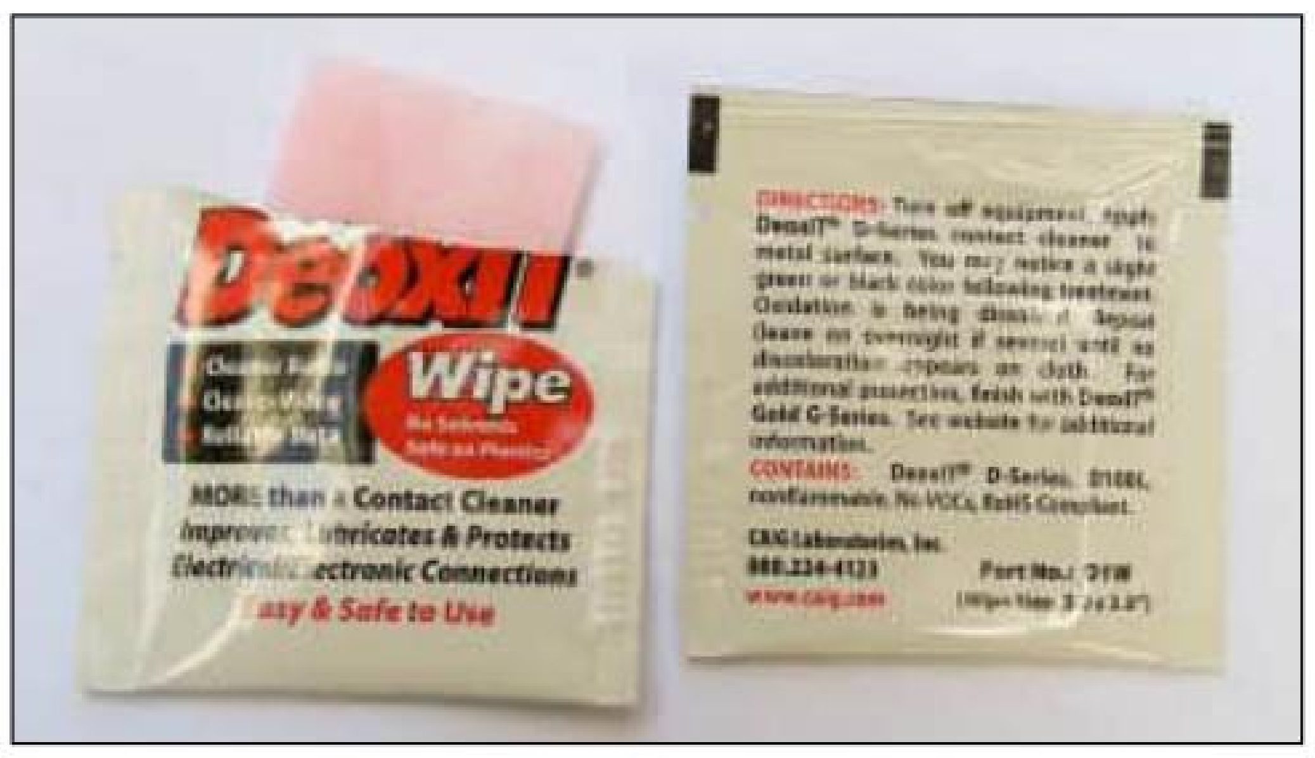

DeoxIT
Wipe
MORE than a Contact Cleaner
Easy & Safe to Use

Fig.20. To get rid of mild oxidation from contacts and wipers these wipes can save the day.

A fine file or emery paper may need to be used. I have very rarely resorted to applying a current-limited high voltage from an insulation tester to clear oxide from sealed wipers, but it can be destructive.

It's essential to apply a bit of fader lube to seal the metal surface once oxides have been cleared; otherwise, oxidation will reoccur. Occasionally, the surface of moulded carbon tracks can become a bit crazed over time and these can be polished down on a glass surface with T-Cut until they are shiny and smooth again. This works wonders with the ModPot carbon tracks once they have been cut out from their plastic frame (Fig.21). The final polish is achieved by rubbing on a flat sheet of paper or cloth.

Solder slaughter

Soldering can kill pots – not immediately – but worse than that, a few months later. The riveted joints from the tags to the silvered track portion are vulnerable to going open circuit due to the corrosive effect of solder flux. I learnt this the hard way when I had this 'great idea' of soldering wires in the rivet holes to adapt wired tag pots into PCB-mount ones for my Colorsound guitar pedals. I had to change a lot of pots under guarantee. Be careful – never allow flux to migrate up the pins to the track.

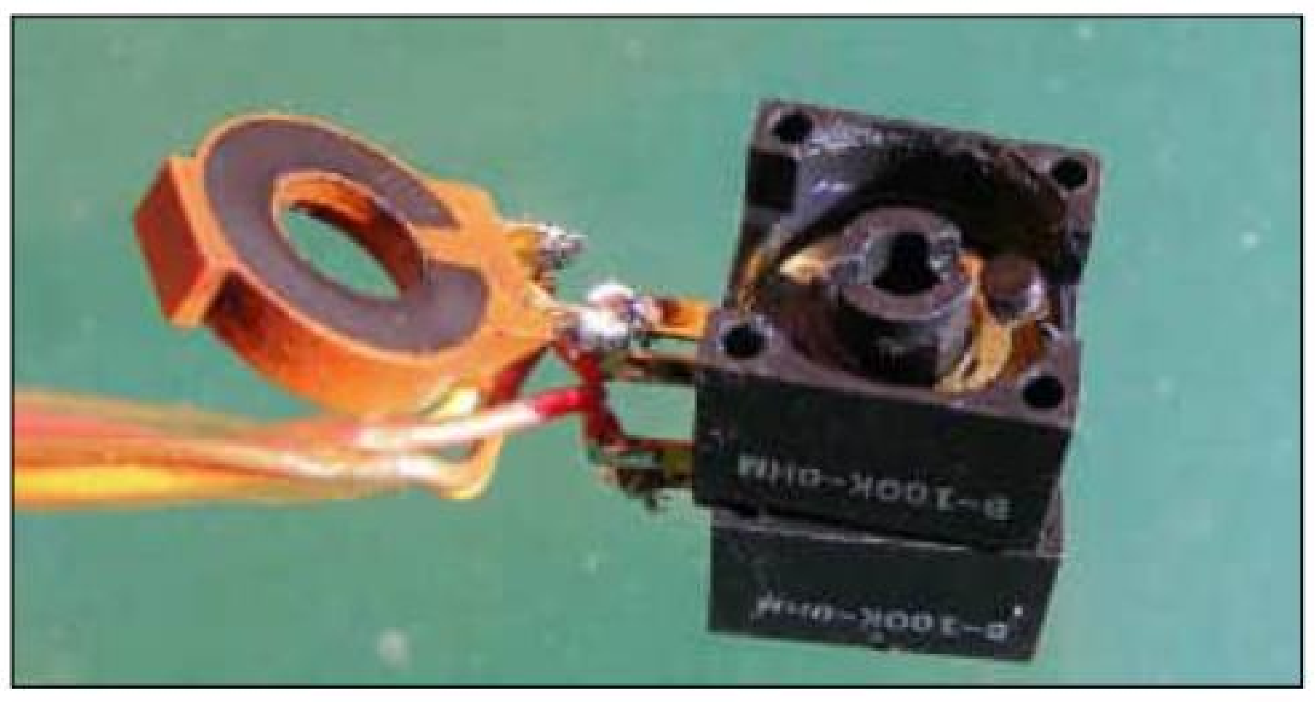


Fig.21. With a bit of cutting and prizing out of its plastic frame the moulded-track element of the ModPot can be extracted for cleaning and polishing. It will then be good for another ten years.

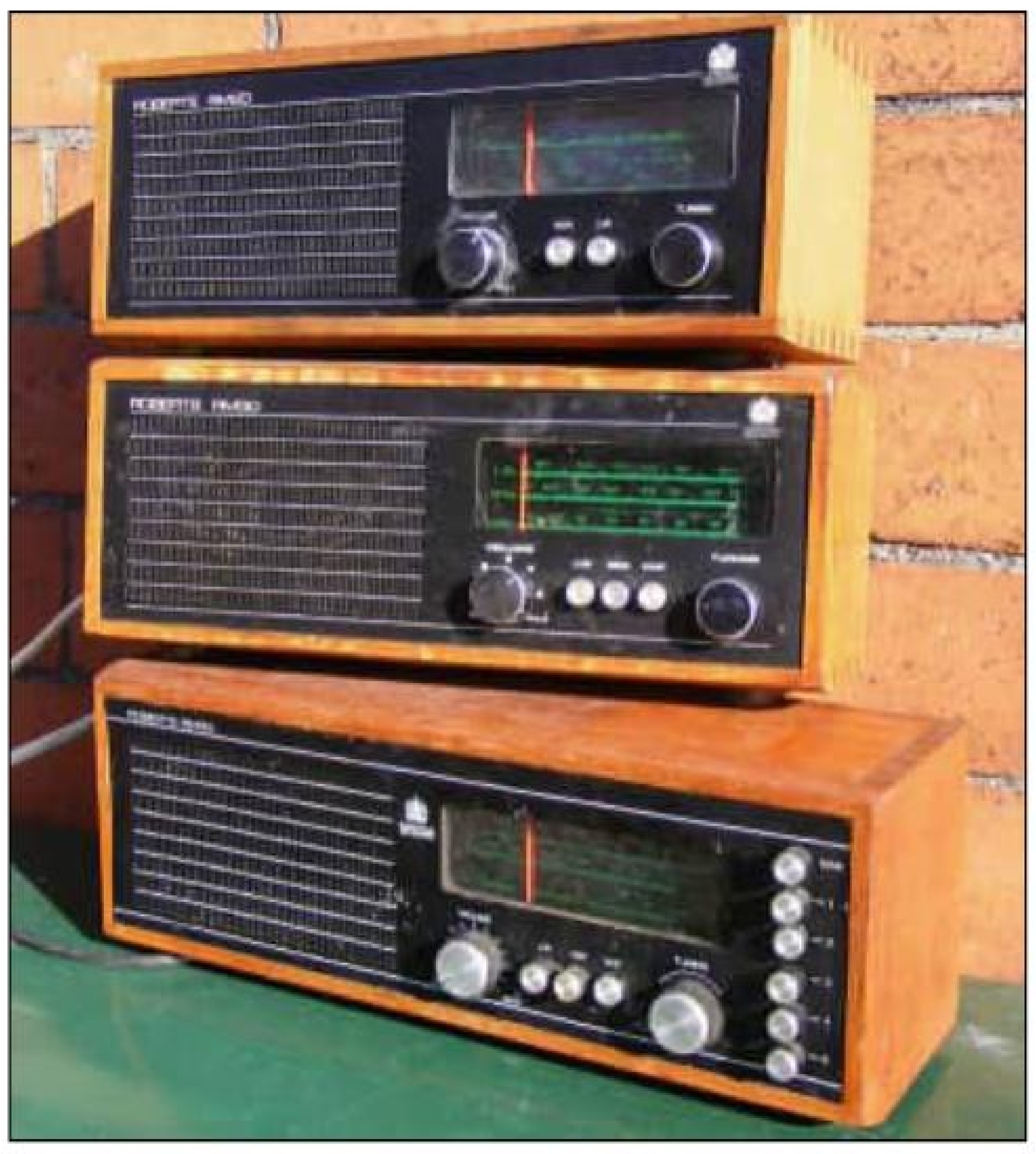


Fig.22. Roberts mains-powered table radios are very popular, but often have bad on/off volume controls that blow your ears off. The earlier RM20, AM-only model is on the top – it does not have this problem since it has a discrete amplifier rather than the high-gain chip used in the two FM models below.

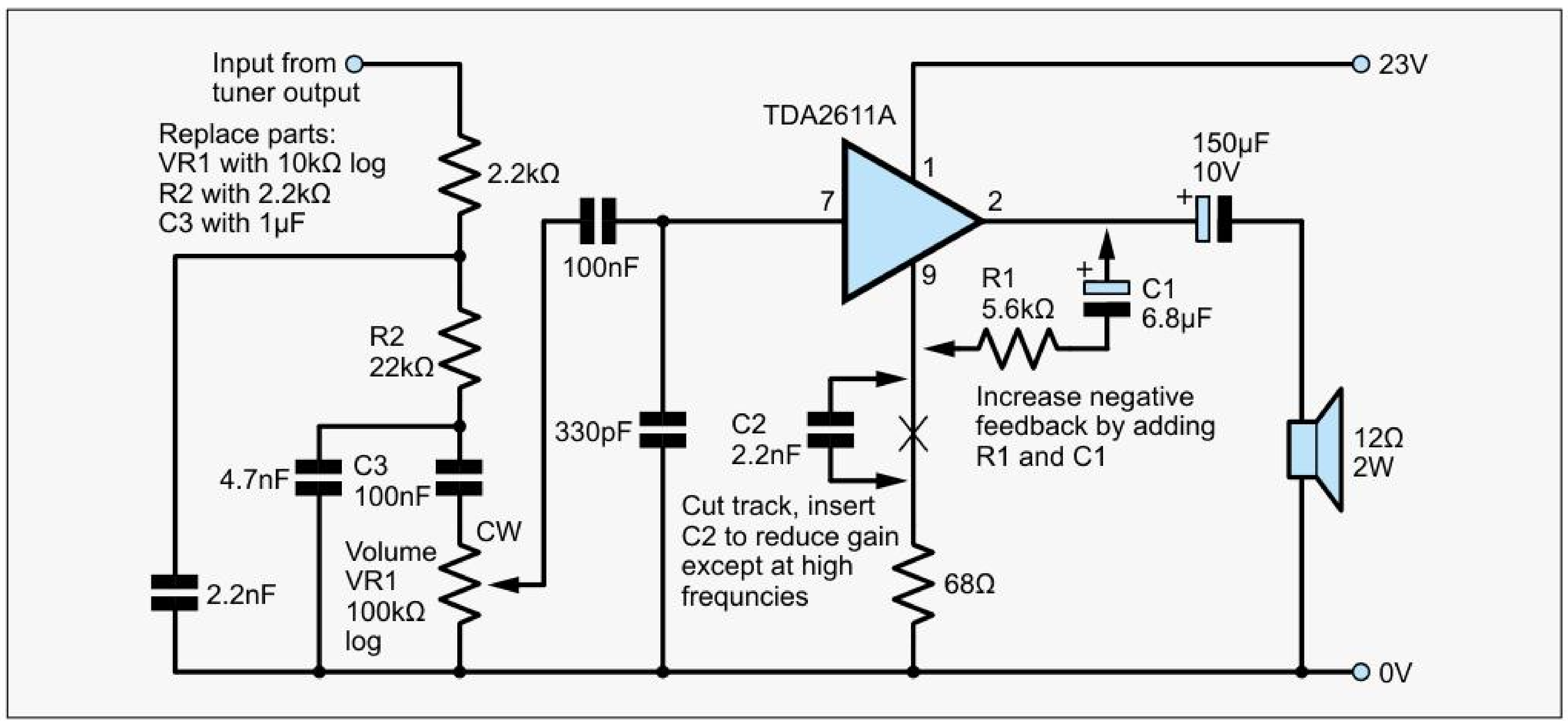


Fig.23. Modifications required to fix the Robert's radio problem. It is quite tricky to reduce the gain of the TDA2611A without it oscillating. The pot value can also be reduced to 10kΩ without distortion.

Switches

As an aside from pots, all the above servicing and DC leakage details also apply to signal switches, another major source of crackles.

Wonky pots

A lot of old radios used combined on/off switch volume controls. With some, the switch was on the mains input, which caused a mild hum. I don't like the idea of mains a few millimetres from the low voltage side of the circuit. (Although, if the metal body of the switched pot unit is connected to mains earth and it passes a 500V insulation test it will be safe.)

Another problem with switched pots is that the volume is turned up before the circuit has time for its DC conditions to settle down and the coupling capacitors to charge up, causing noise as currents are flowing through the pot. One possible solution is to use a push/pull switch, assuming the volume level is at the previous level. Another fix is a time-delay output mute circuit. Best of all would be to use a separate switch and volume control. The Robert's mains table radios shown in Fig.22 have integral mains switches and over 40dB of gain after the volume control. There's practically an audio 'explosion' when you turn them on. This can be fixed by reducing the pot value from 100kΩ to around 5-22kΩ, increasing the input coupling capacitor and reducing the amplifier gain with additional negative feedback components around the chip. This is shown in Fig.23. These radios have exposed metalwork screws and only two core mains leads. To make them safe and compliant they should be earthed and upgraded to a three-core lead.

Going over the edge

One of my least favourite types of pot is the 'edgewise' design – intended for thumb operation. When they fail it's often impossible to find a replacement, so I usually give up and try to fit a rotary device. I recently spent six months struggling to source an edgewise switched audio-taper pot when designing the latest Dubreq synthesiser because the 'style concept' demanded a thumbwheel. I tried eight mostly useless samples, and when the first units came off the line, the knobs were rubbing against the slot, needing more rework (see Fig.24). Avoid these nuisances in your designs!

Sliders

I also dislike slider pots because they're a pig to fit and dirt falls into the open slot. Also, they come in 20 different sizes with multiple non-standard knobs that easily get lost. On my 'can't be bothered to repair' rack a lot of the equipment

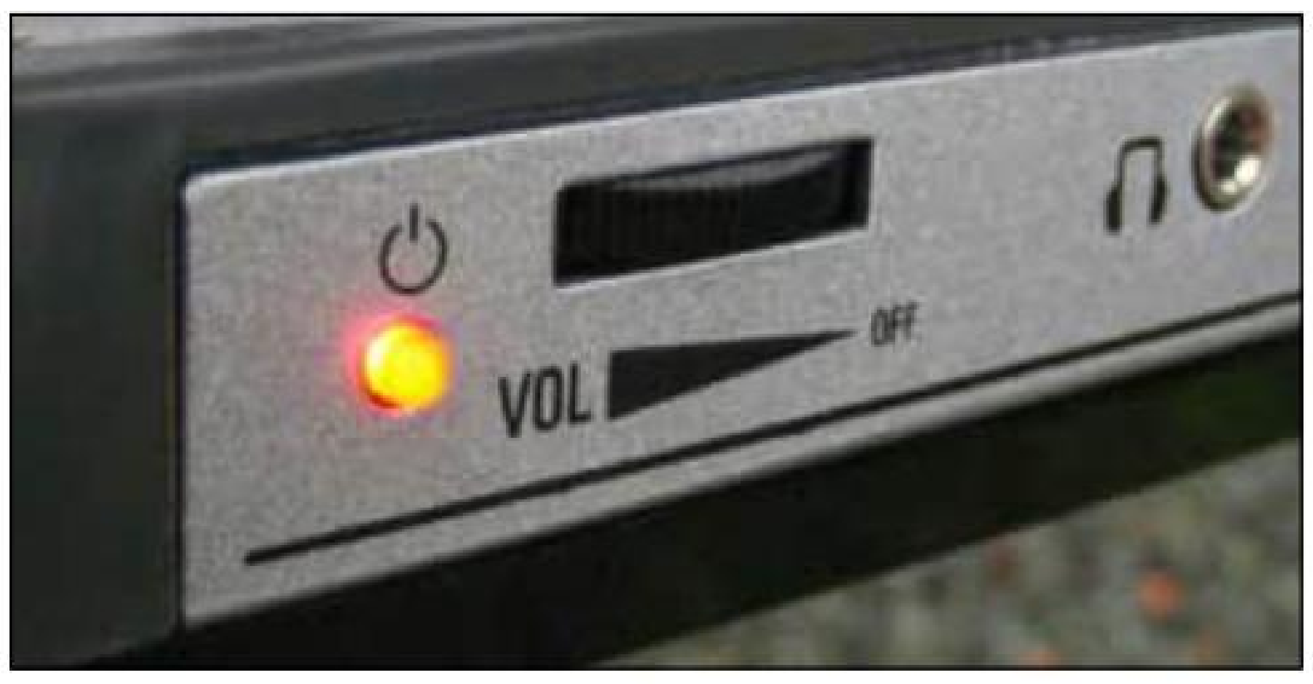


Fig.24. This little edgewise switched pot on my latest synth design caused a few months of headaches. In your designs, use a sensible rotary pot and a latching power switch.

Fig.25. A nice synth ruined by stuck sliders used only for 'style' purposes. Fixing these will be a £200 job. In this case it's worth doing because the machine is a highly collectable ARP.

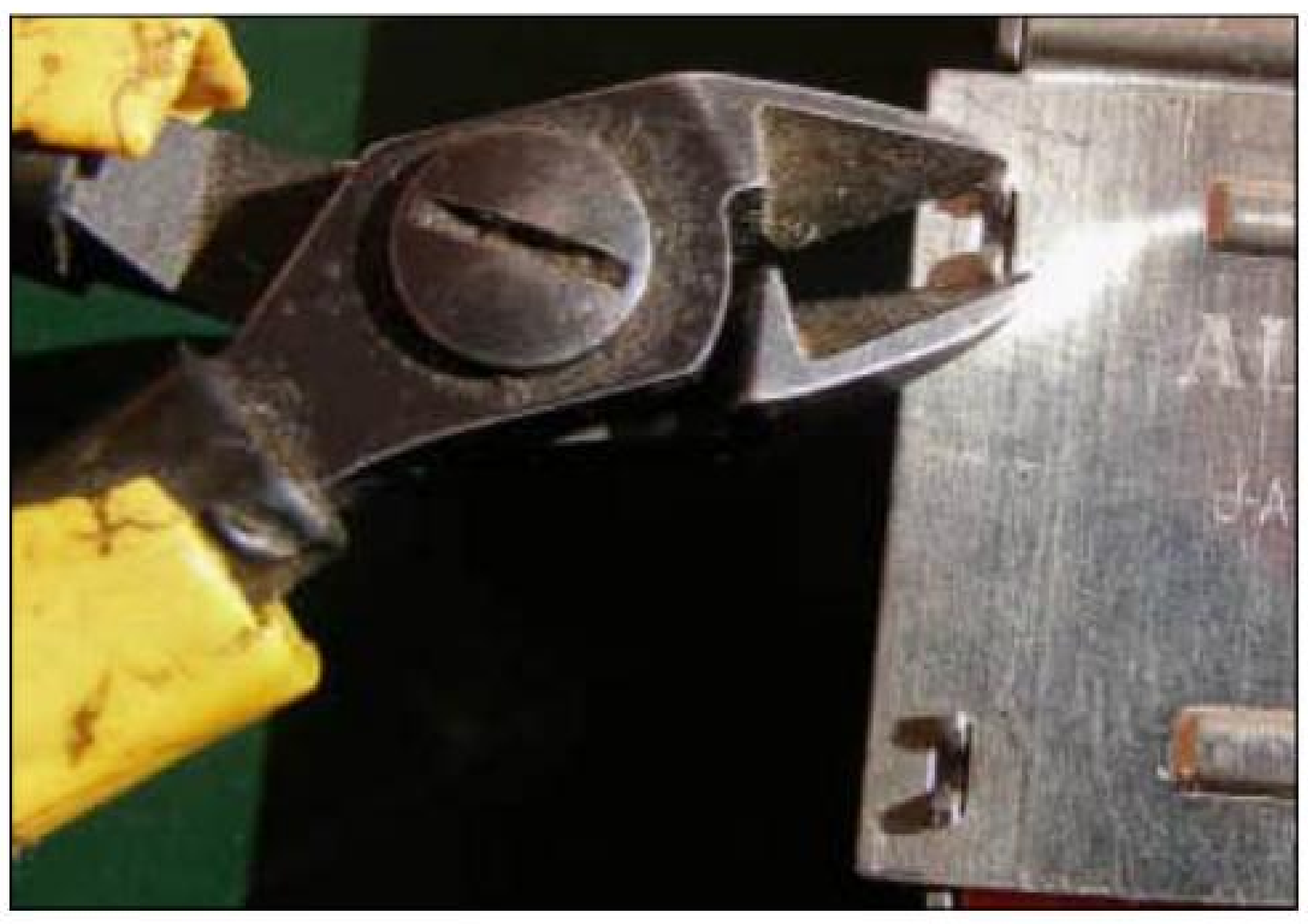


Fig.26. Opening up an Alps mixer fader for cleaning. Squeeze the tabs to get the side plate off.

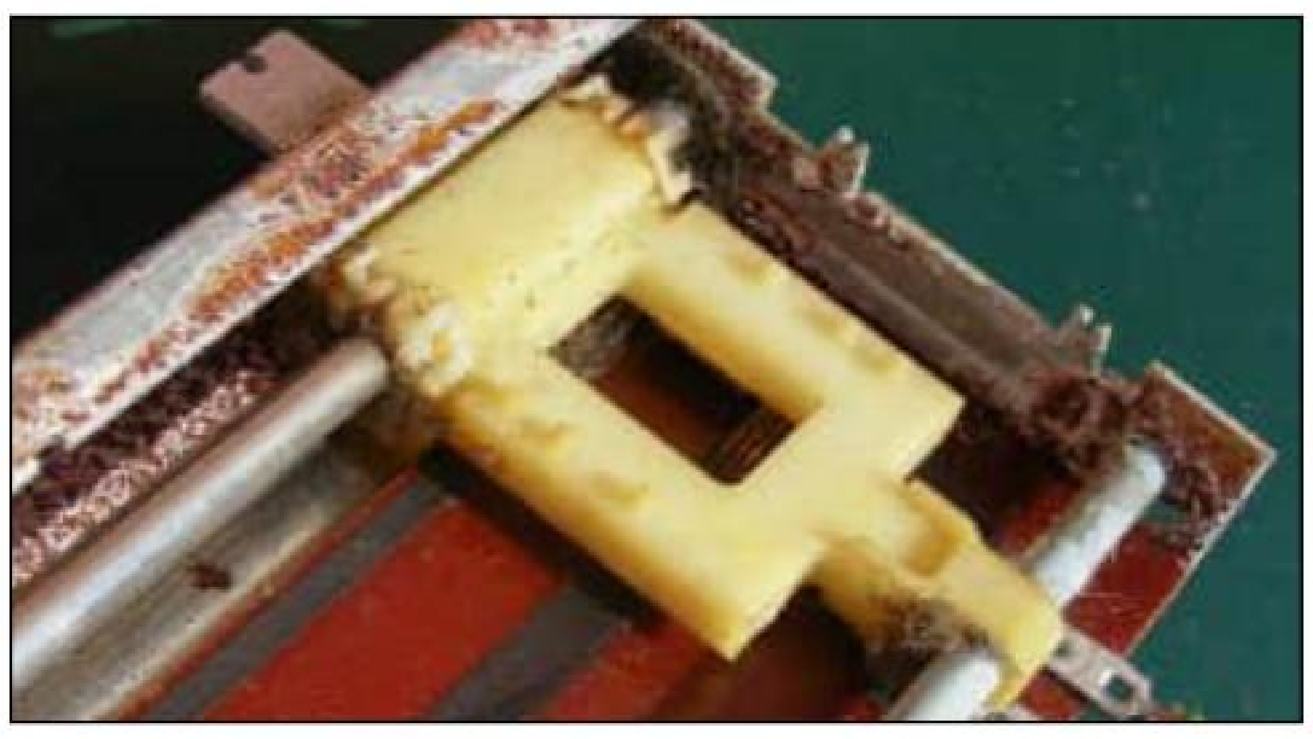


Fig.27. Typical fader rubbish to be cleaned out.

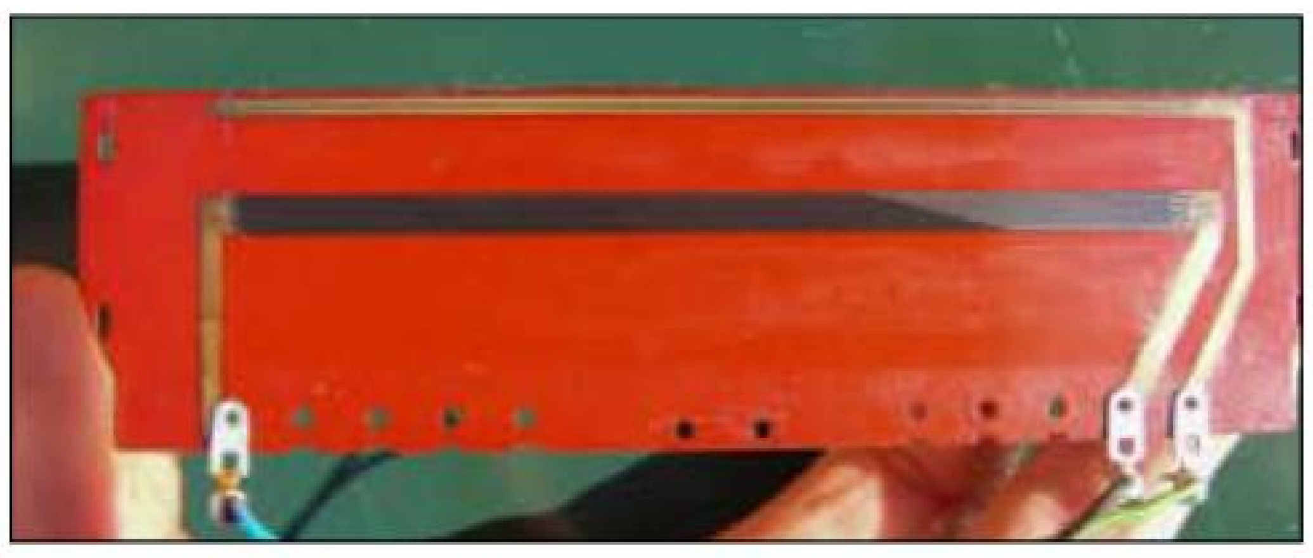


Fig.28. A nicely cleaned track ready for wiper de-oxidation and application of lubricants.

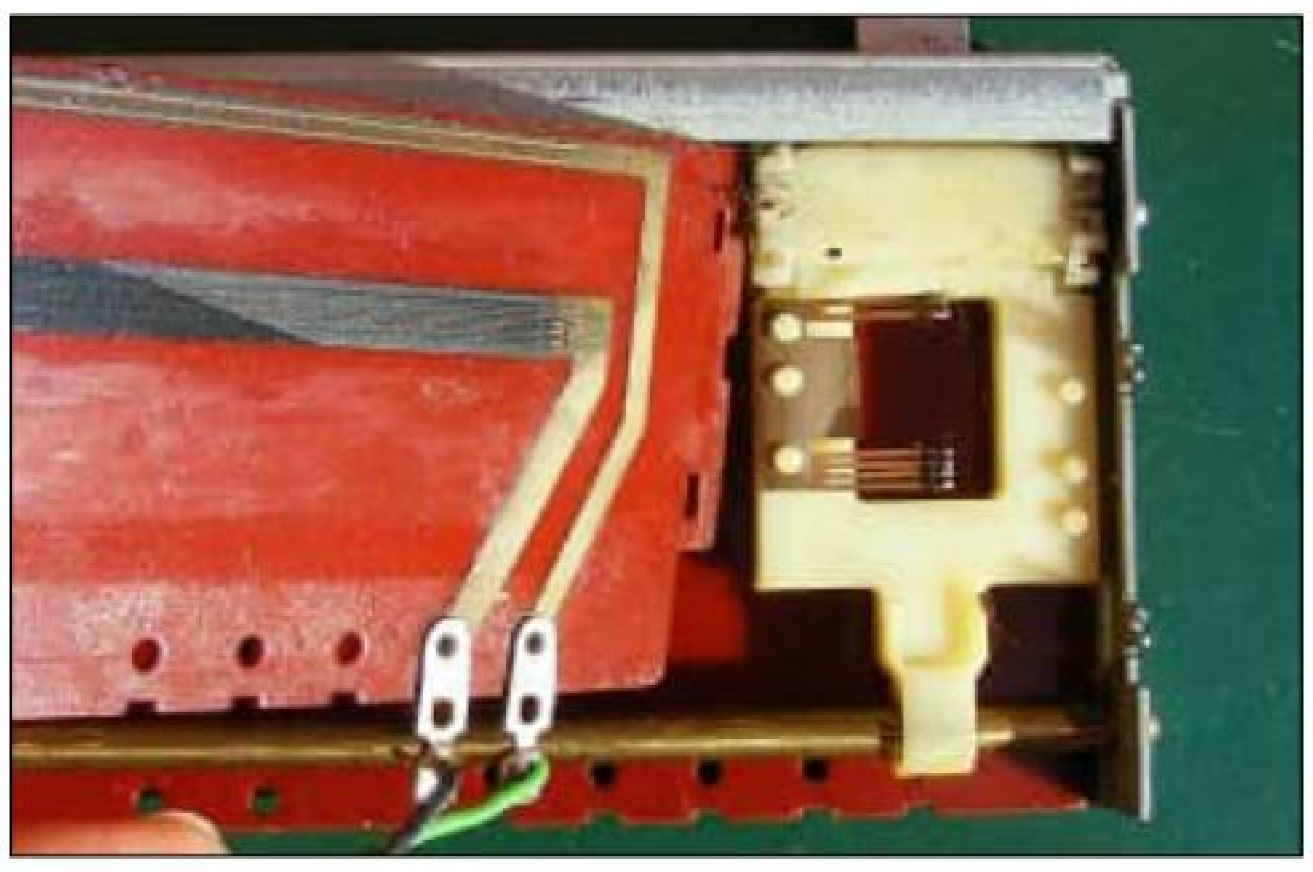


Fig.29. A small sable-hair painter's brush was used to apply the track lubricant.

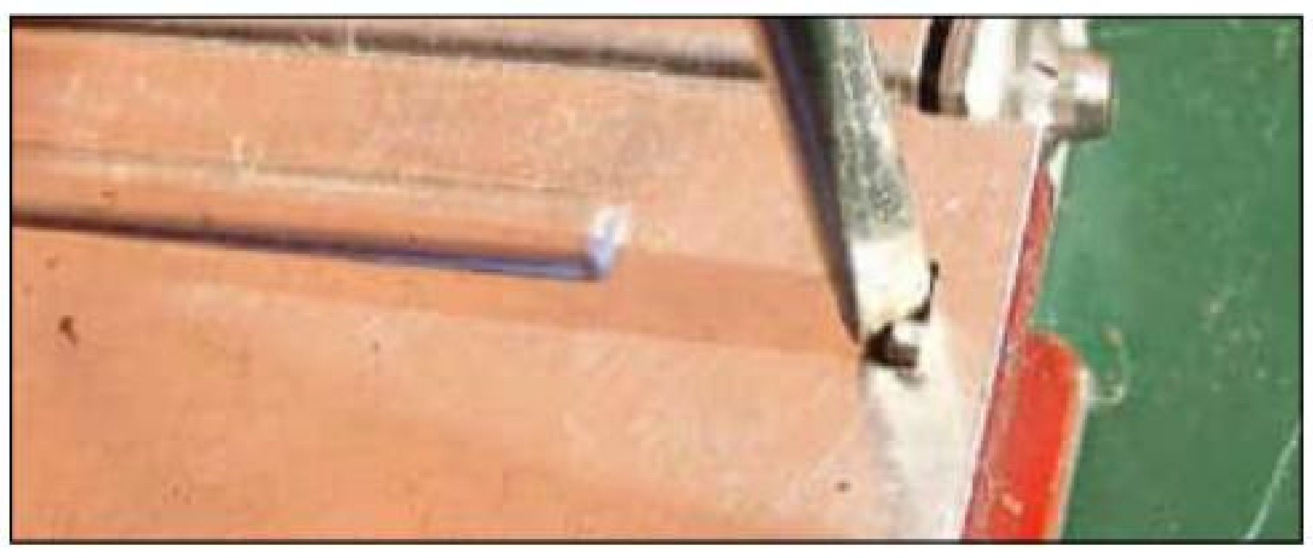


Fig.30. Re-crimping the tabs by twisting with a screwdriver.

Fig.31. Bush TR130 with a Plessey switched volume-control pot.

needs new unobtainable sliders like the synthesiser shown in Fig.25. As an engineer, I want to avoid odd-shaped pots, industrial designers please take note. I only make an exception for mixer faders because they are a cultural norm.

The Alps fader was popular on older semi-pro mixing desks. Like all slider pots they suffer from debris entering the slot. Fig.26 to Fig.30 shows a sequence of photos detailing how to recondition them.

Potting bushes

In some cases, just replacing the input electrolytic capacitor on the wiper with a polyester of lower value solves scratching. I do this in the popular Bush TR130 radio circuit shown in Fig.31. I find a Mullard 1μF C280 'tropical fish' capacitor looks right in a 1967 radio and fits in the PCB nicely. The resulting higher bass cut-off frequency is of no consequence given the cut-off frequency of the speaker and output transformer. The circuit is shown in Fig.32.

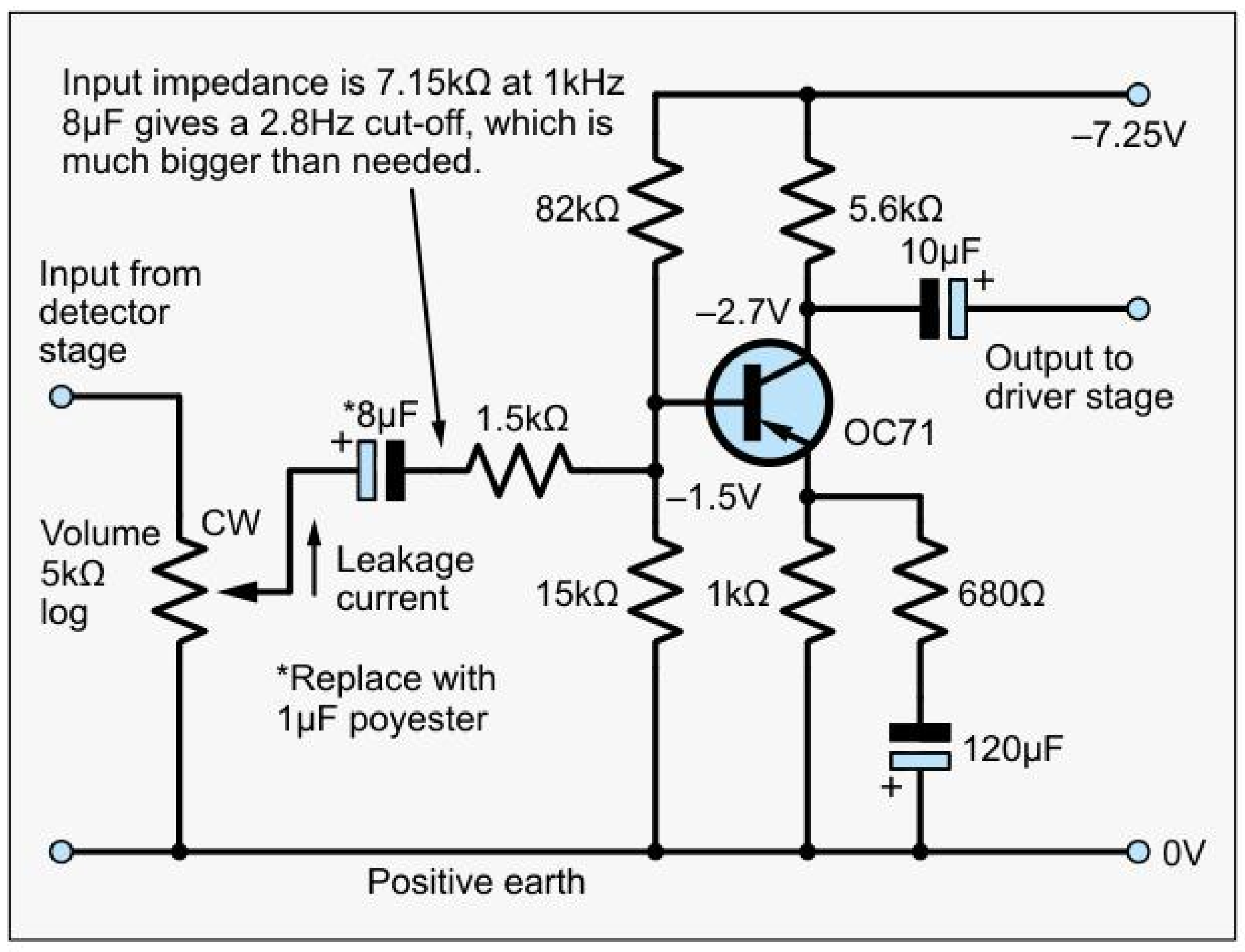

Input impedance is 7.15kΩ at 1kHz
8μF gives a 2.8Hz cut-off, which is much bigger than needed.
–7.25V
82kΩ
5.6kΩ
10μF
+
Input from detector stage
–2.7V
Output to driver stage
*8μF
+
1.5kΩ
OC71
–1.5V
Volume 5kΩ log
CW
Leakage current
15kΩ
1kΩ
680Ω
*Replace with 1μF poyester
120μF
+
Positive earth
0V

Fig.32. The volume control and post-fader amplifier circuit on the Bush TR130. Note that the wiper coupling capacitor should be replaced with a lower capacitance plastic-film unit to reduce leakage and settling time.

A common problem with old AM radios is the use of the volume control track element as the diode detector load. This can put over a volt of DC across the track. The solution is to provide a separate resistor for this. This should be coupled via a capacitor to a high-impedance load at least 4x higher. This is needed because distortion can arise if the load seen by the diode is different for AC and DC – see Fig.33.

In some circuits, pots are placed in negative feedback loops such as in Fig.34. These active volume controls are extremely sensitive to pot noise because if the wiper loses contact it goes to maximum volume rather than minimum. In some circuits the output can even go hard offset if the designer is daft enough to route the negative feedback DC path through the pot. I had endless trouble with an Omeg volume pot in the Douglas Self *Wireless World* 1979 pre-amp even though it was properly coupled. This was transformed when I replaced it with a Bourns 91 conductive plastic unit coupled with metal-cased ITT TAA tantalum capacitors. These had 10x less leakage current than the Mullard 017 blue electrolytic capacitors I used originally.

Settling time

Equipment that has not been switched on for a long time may suffer leakage currents and terrible pot noise while wet electrolytic capacitors reform. This is why I have always advocated tantalum coupling capacitors in my designs (along with the necessary distortion-mitigation measures). There is also the normal time all capacitors take to charge up when turned on. In some designs – using high value pots (around 100kΩ) and large value (around 100μF) capacitors – settling time can take a few minutes. This is usually the result of a misguided attempt to extend bass and obtain ultra-low distortion down

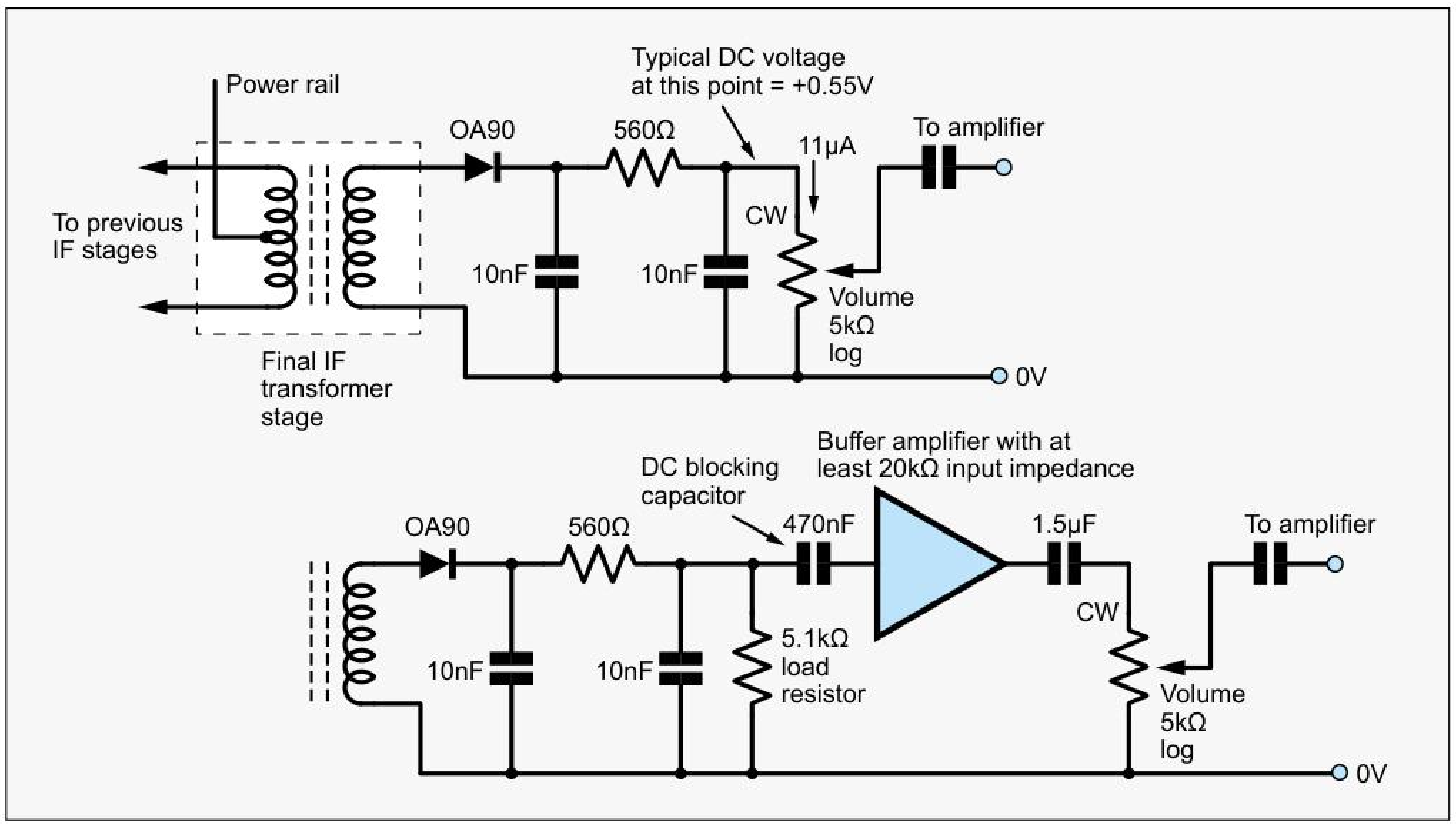
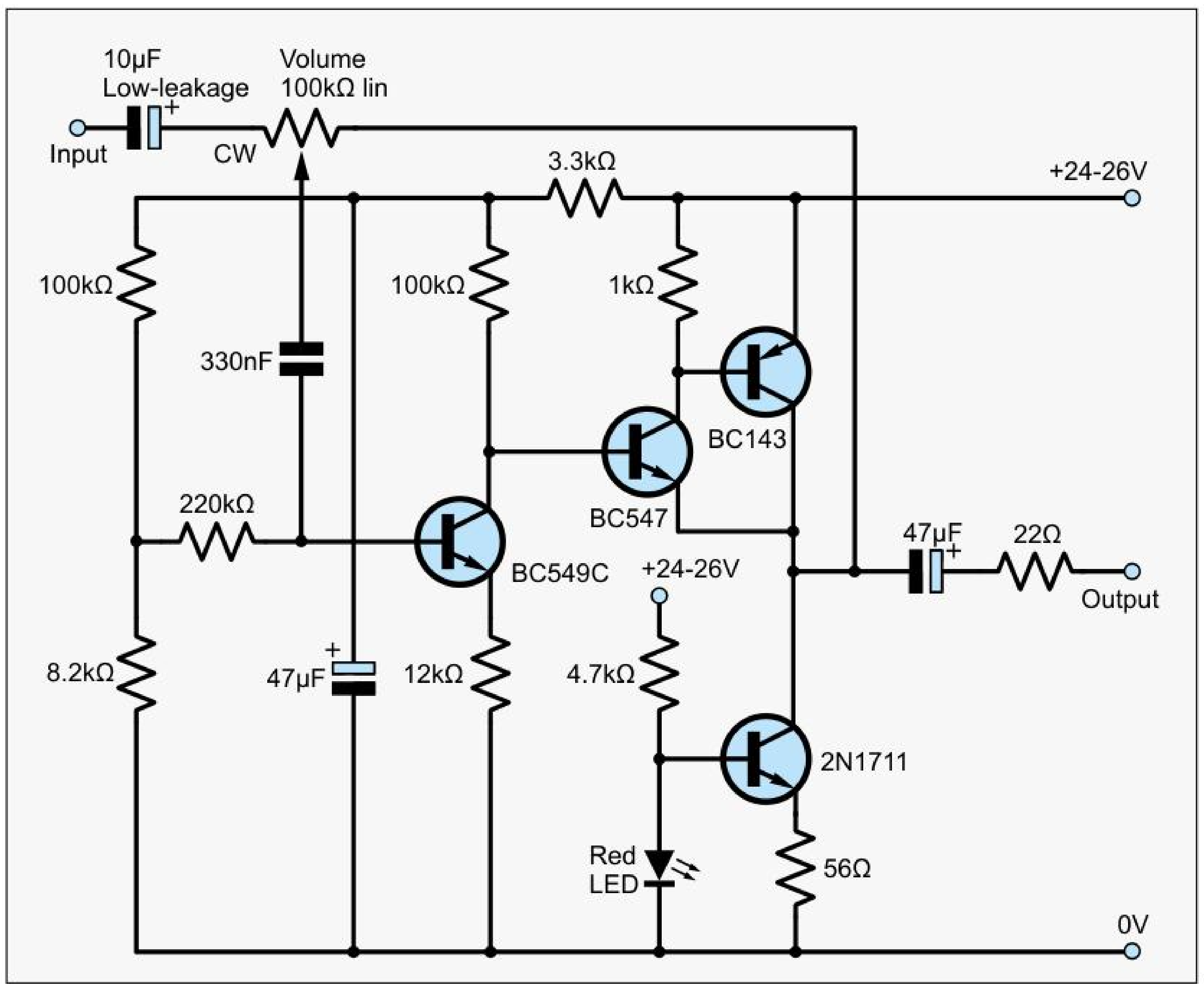

Power rail
Typical DC voltage at this point = +0.55V
OA90
560Ω
11μA
To amplifier
To previous IF stages
CW
10nF
10nF
Volume 5kΩ log
0V
Final IF transformer stage
Buffer amplifier with at least 20kΩ input impedance
DC blocking capacitor
OA90
560Ω
470nF
1.5μF
To amplifier
CW
5.1kΩ load resistor
10nF
10nF
Volume 5kΩ log
0V

Fig.33. A lesser known cause of pot scratching on many old radios, including the TR130, is the practice of using the volume pot track as the detector load.



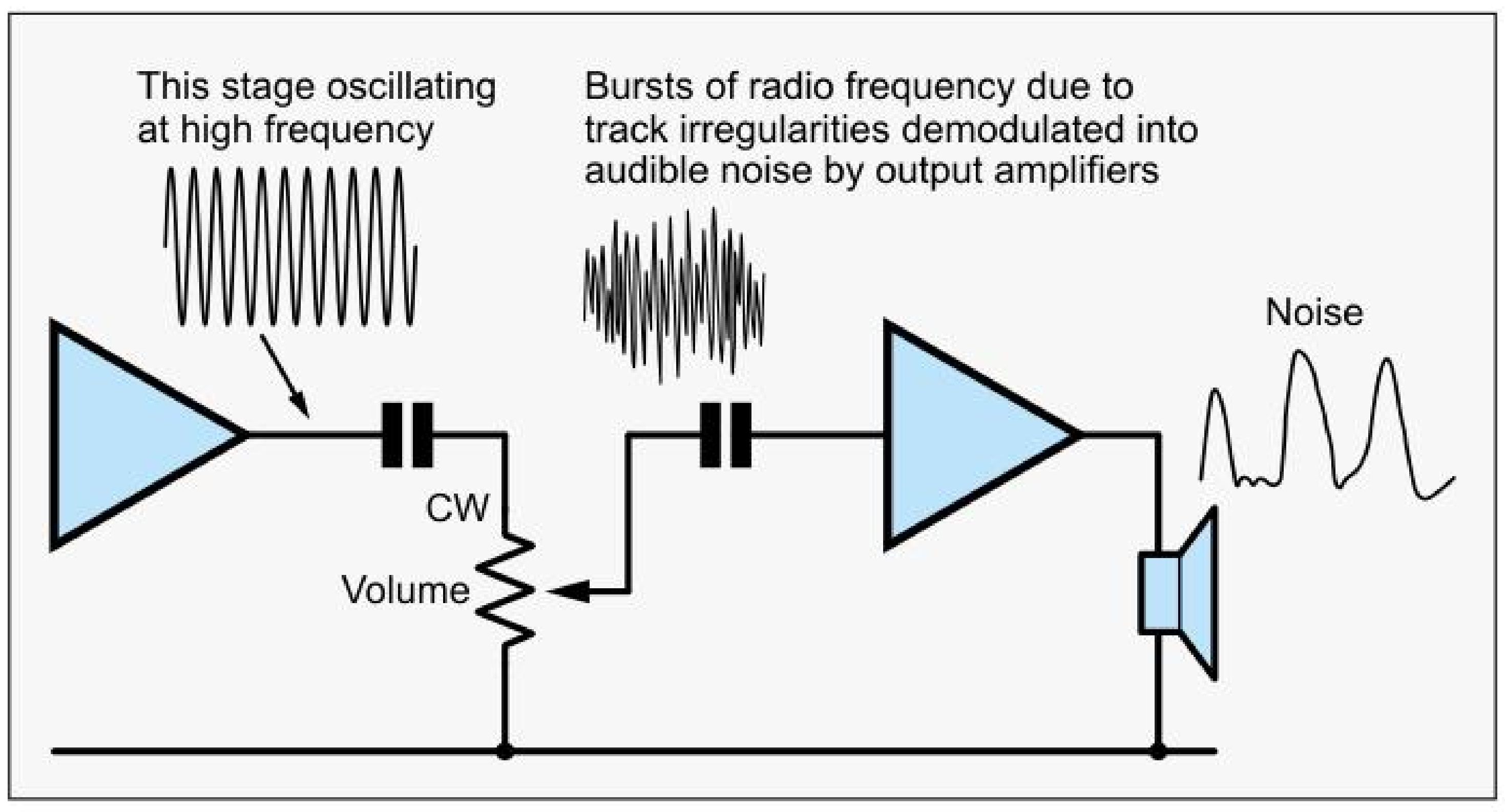

This stage oscillating
at high frequency
Bursts of radio frequency due to
track irregularities demodulated into
audible noise by output amplifiers
Noise
CW
Volume

Fig.35. (above) Occasionally, radio-frequency oscillation or breakthrough from inadequate filtering may give pot crackling similar to DC on the track.

Fig.34. (left) When a pot is used as a negative feedback volume control element, such as in this Baxandall topology circuit, a jump to high resistance caused by dirt on the track can cause the gain to go open loop with horrific noise. Where necessary, arrange for the feedback to be AC-coupled through the pot with a separate path for DC.

to subsonics. Recording engineers are well aware of these settling phenomena and usually allow their equipment to stabilise before use.

Use it or lose it

Pots must be rotated over their full rotation every six months or so, otherwise they may seize up or oxidise – just like a car or any other machine that relies on lubrication, they have to be used.

High-frequency oscillations

If a pot is fed with an ultrasonic high frequency signal, for example from an oscillating op amp, it can be demodulated later in a following amplifier (Fig.35). The result sounds just like DC leakage crackling and it's caught me out a few times. This is a relatively rare pehenomenon, but is worth knowing about if apparent pot crackle is resisting all attempts at being cured.

Pot pourri

In this section we'll list a selection of pots with their expected rotational life (RL) specification before they wear out. Those with a life of over 20,000 turns are expensive (£4-12) compared to the £0.70 to £2.00 for 'regular' carbon track units. This means 'life cost' should be considered. There's no point using a high-spec pot for a rarely used treble control on a Hi-Fi amplifier, but it's a good idea for the volume control, since a £50.00 service fee could be avoided later. A lot of these pots are no longer made and there have been many mergers and acquisitions in the components industry. This means old designs often turn up again under a different name. They are included because they are regularly found in old equipment and as NOS 'new old stock' on auction websites. Plus, I've taken all of these apart for analysis so you don't have to!

Allen-Bradley series 70 ModPots (100,000 RL)
Any configuration possible, easy to disassemble. Poor humidity resistance. conductive plastic, cermet and moulded track versions.

Allen-Bradley J series moulded track (100,000 RL – Extra life (EJ) version 1,000,000 RL)
Very strong, sealed, large 2W mil spec. Used in 'Cry Baby wah-wahs' effects pedals. Been in production for over 60 years. Now made by Honeywell. Stainless-steel case. Similar types made by other manufacturers as the military RV4 form. Clarostat make a conductive plastic version called the 380 series. See Fig.36.

Allen-Bradley W series 10mm sealed miniature moulded-track (50,000 RL)
Very solid design. See Fig.37.

Alpha (16,000 RL)
The best cheap pots, a Taiwanese version of the Japanese Alps pots. I use them in my consumer products. They use four silvered wipers on track, three on the slip ring.

Alps (16,000 RL)
Good for most Hi-Fi – the 27mm Blue Velvet is a definitive Hi-Fi volume control.

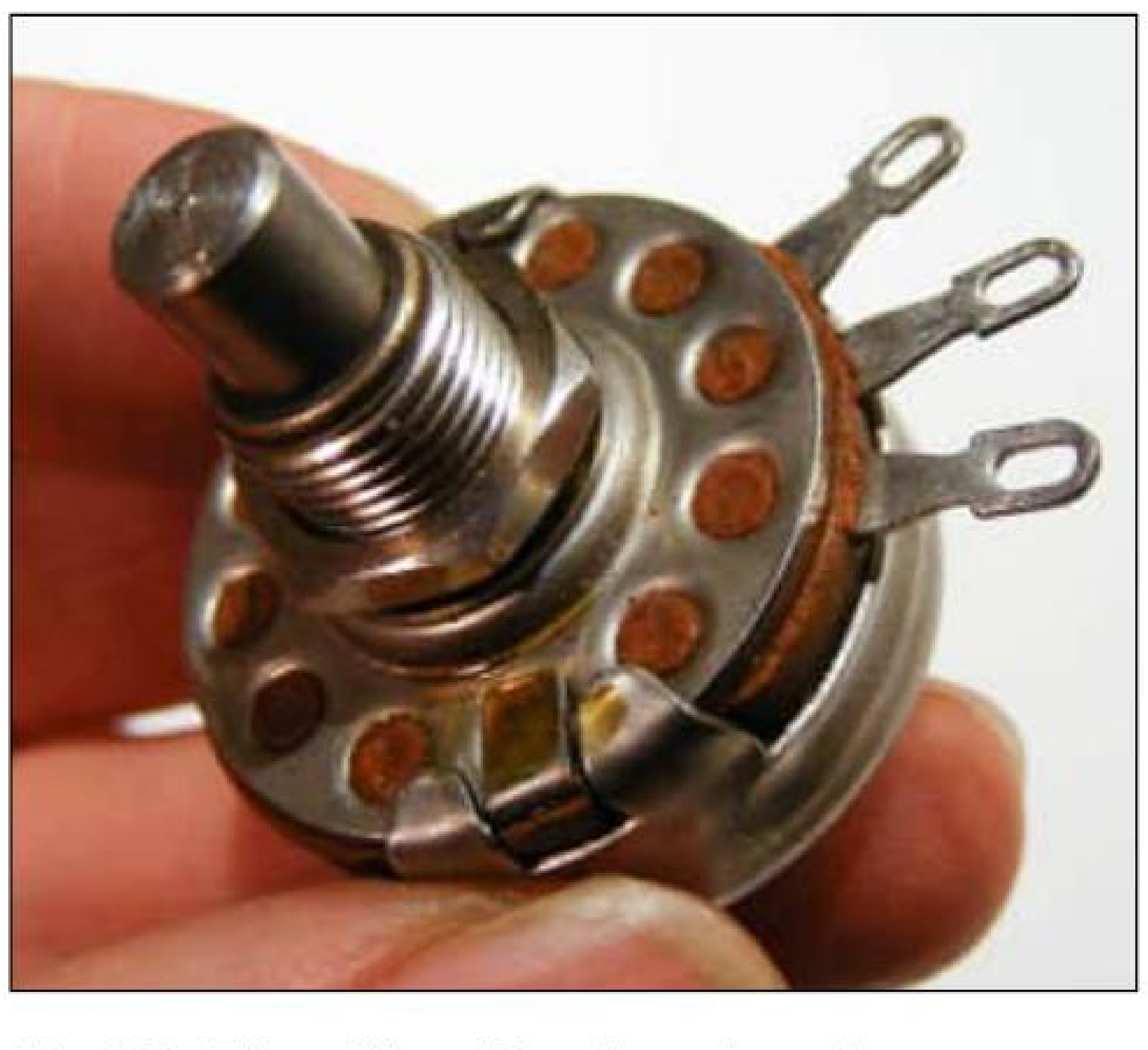


Fig.36. The Allen-Bradley J-series; one of the toughest pots made and with low rotational noise due to a highly polished moulded-track element. I've never had to replace them, only scavenged them from old equipment for reuse.

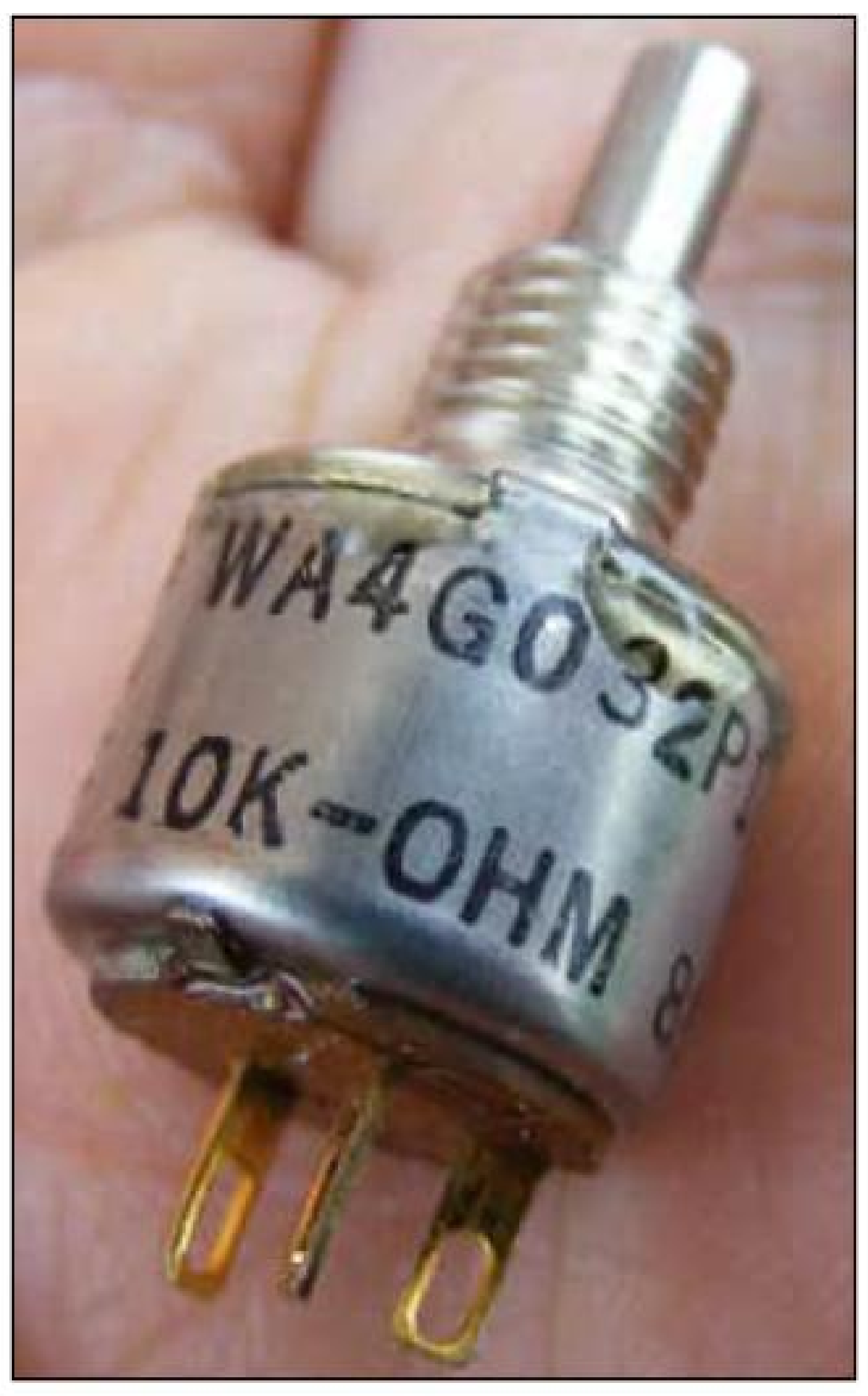


Fig.37. The Allen-Bradley W-series, a quiet sealed pot that's very small.

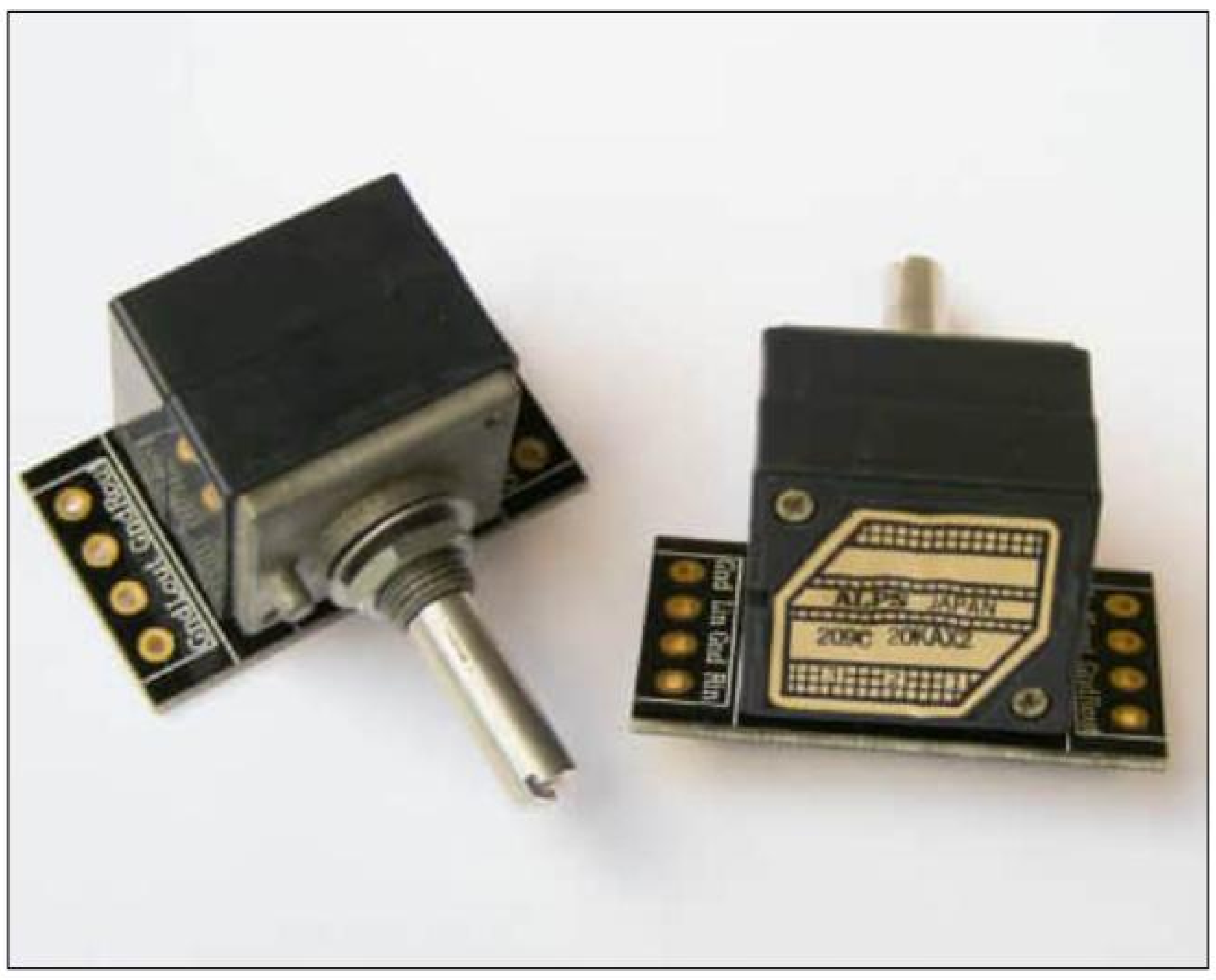


Fig.38. A pot loved by Hi-Fi amp designers, the ALPs Blue Velvet – the smoothest carbon-track pot made. Available from Mouser and eBay.

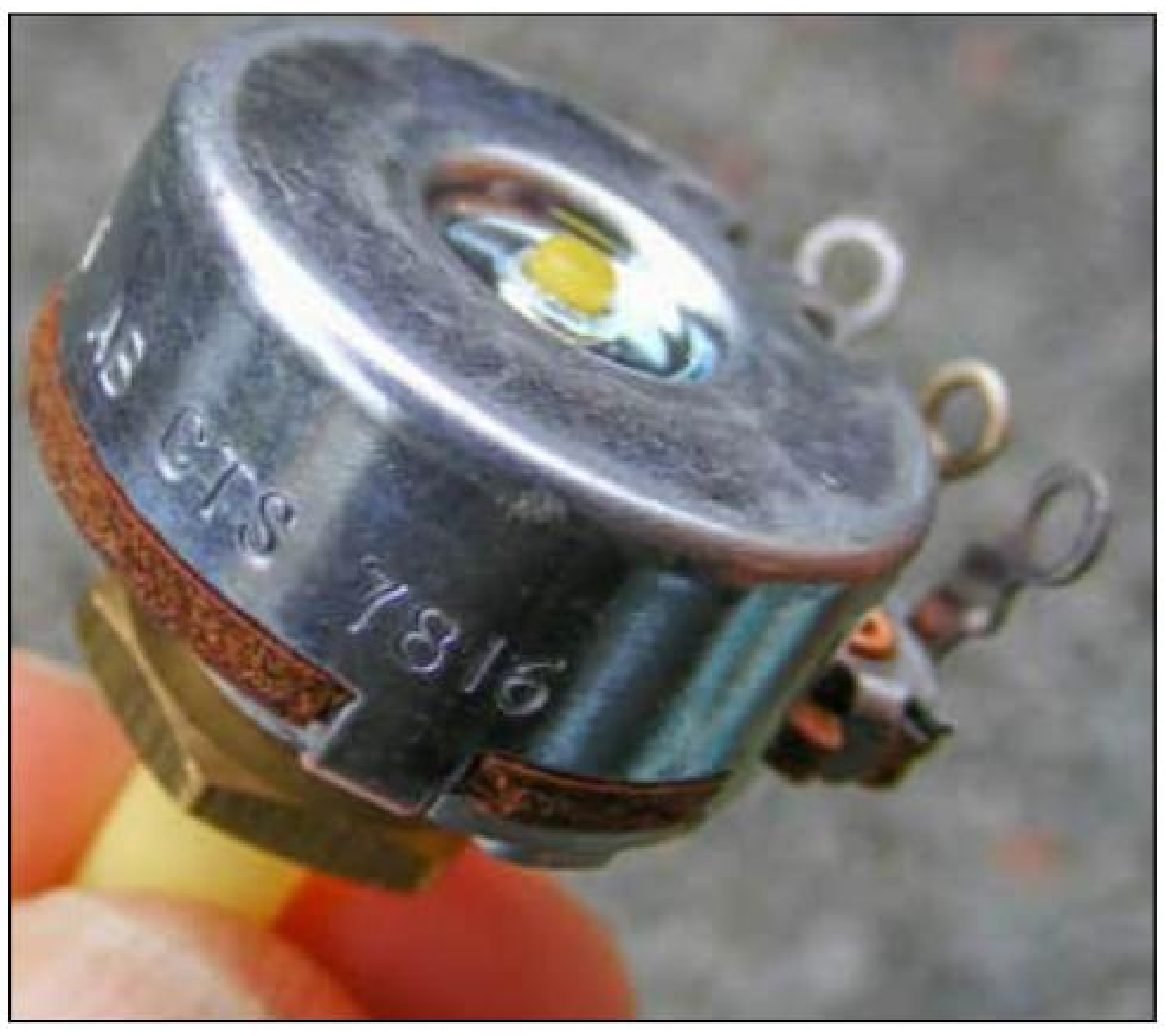


Fig.39. AB (not to be confused with Allen-Bradley) CTS pot. Tough old-fashioned carbon track product made in Wales by Blore Edwards Ltd. Easily identified by the plastic spindle pip at the back.

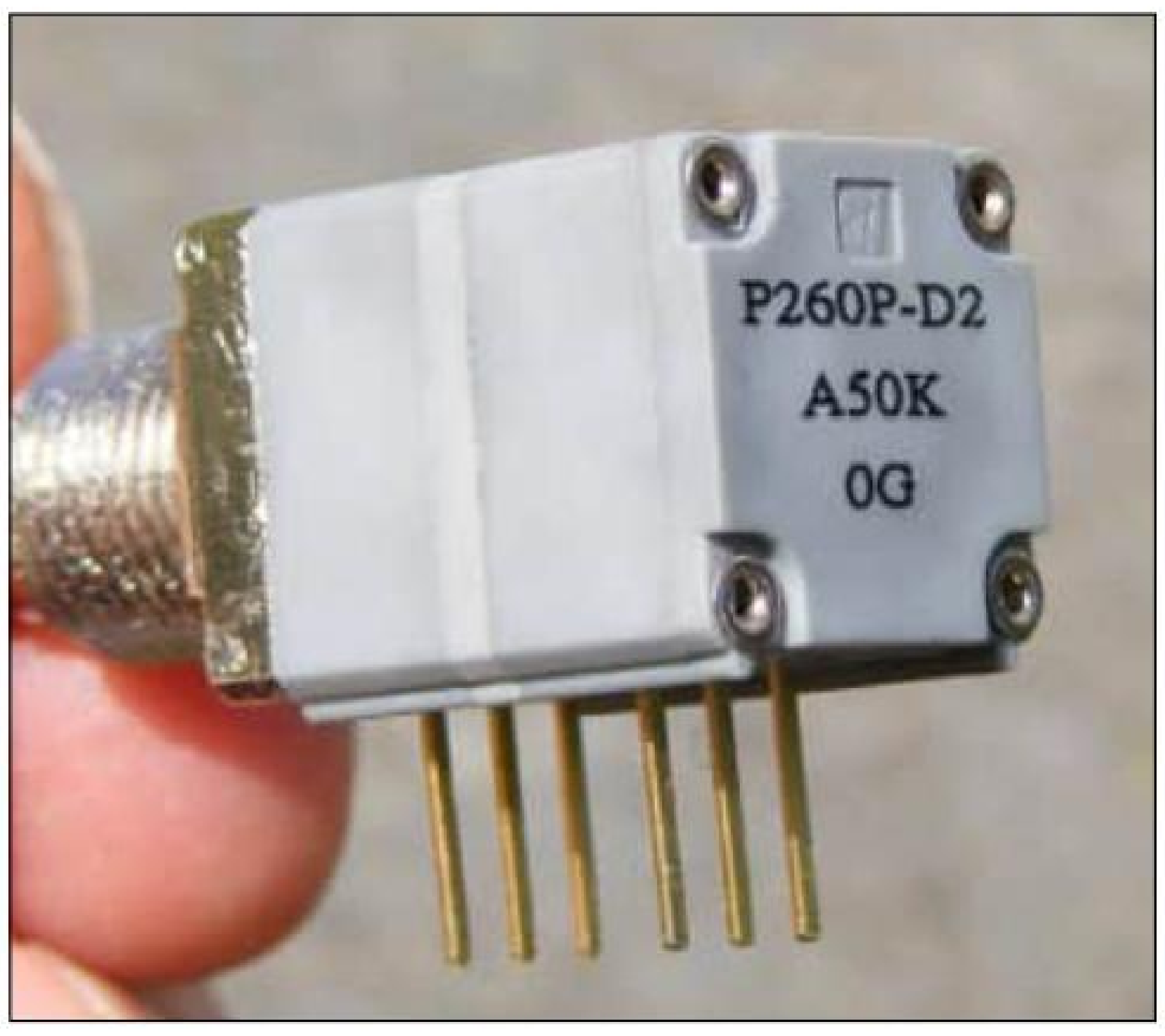


Fig.40. BI P260 – the most cost-effective sealed conductive-plastic pot. This device is very small, good for PCBs, uses a four-fingered wiper and is distributed by Profusion.

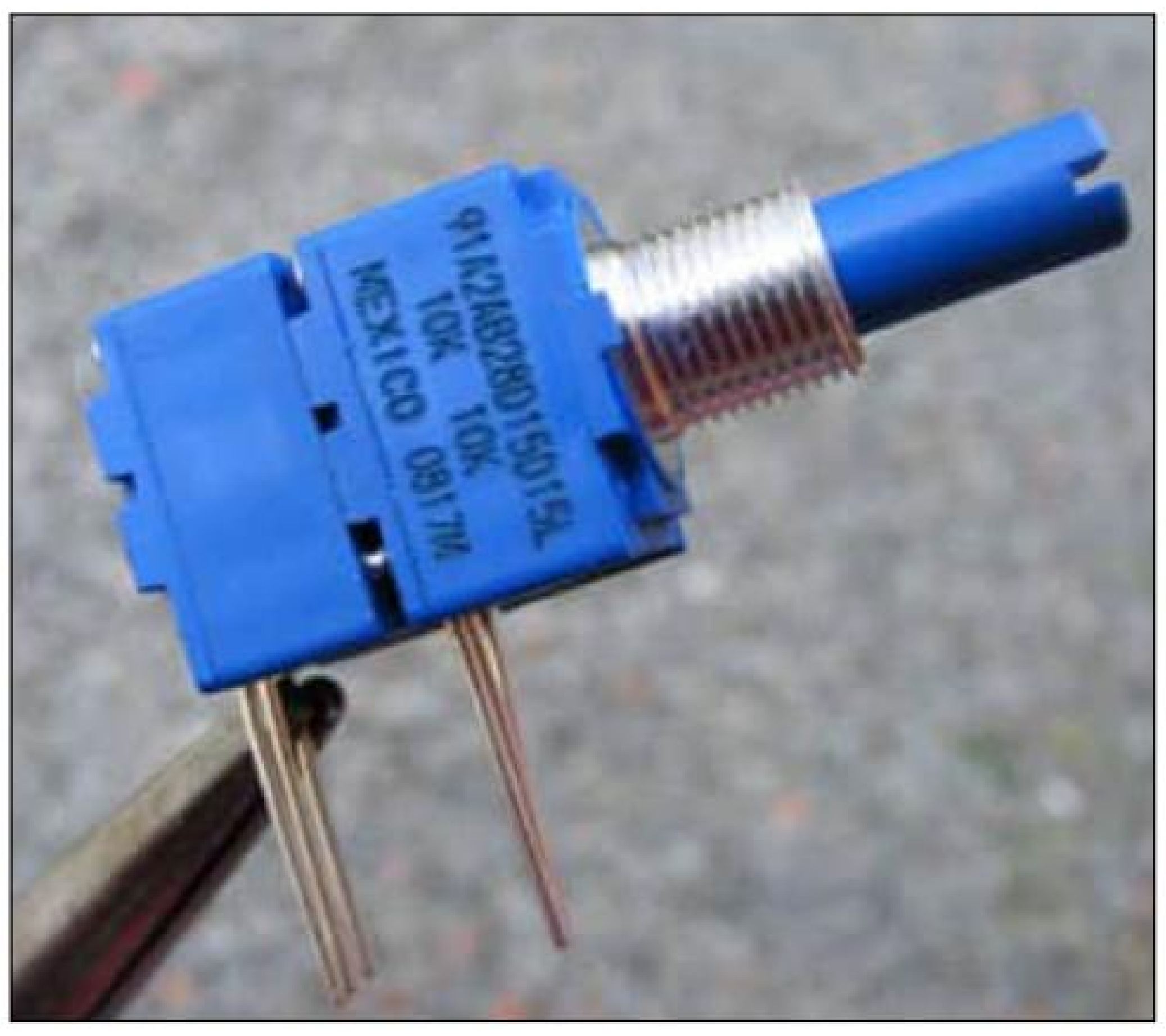


Fig.41. The Bourns 91 is my favourite volume control. Its plastic construction is a bit weak, but its noise performance is the best – available from Farnell and Audio Out.

Excellent track matching, especially at low levels. See Fig.38.

AB CTS series (now Blore Edwards) (25,000 RL)
Physically strong, used in EMS 4000 series synths and the Leak Delta 70 amplifier. See Fig.39.

BI P260 series (1,000,000 RL – this seems optimistic!)
High initial jump-on step resistance, good price for a sealed conductive plastic pot. See Fig.40.

Bourns 51, 81, 82, 91, 95, 96 series (100,000 RL)
My favourite, even the cermet versions (50,000 rotations) are quiet. Used in professional broadcast audio. At around £12 from Farnell (order code 9357858) for a stereo 10kΩ log 91A2A-B28-D15/D15L they are expensive, but worth it. Unique blue appearance, as shown in Fig.41.

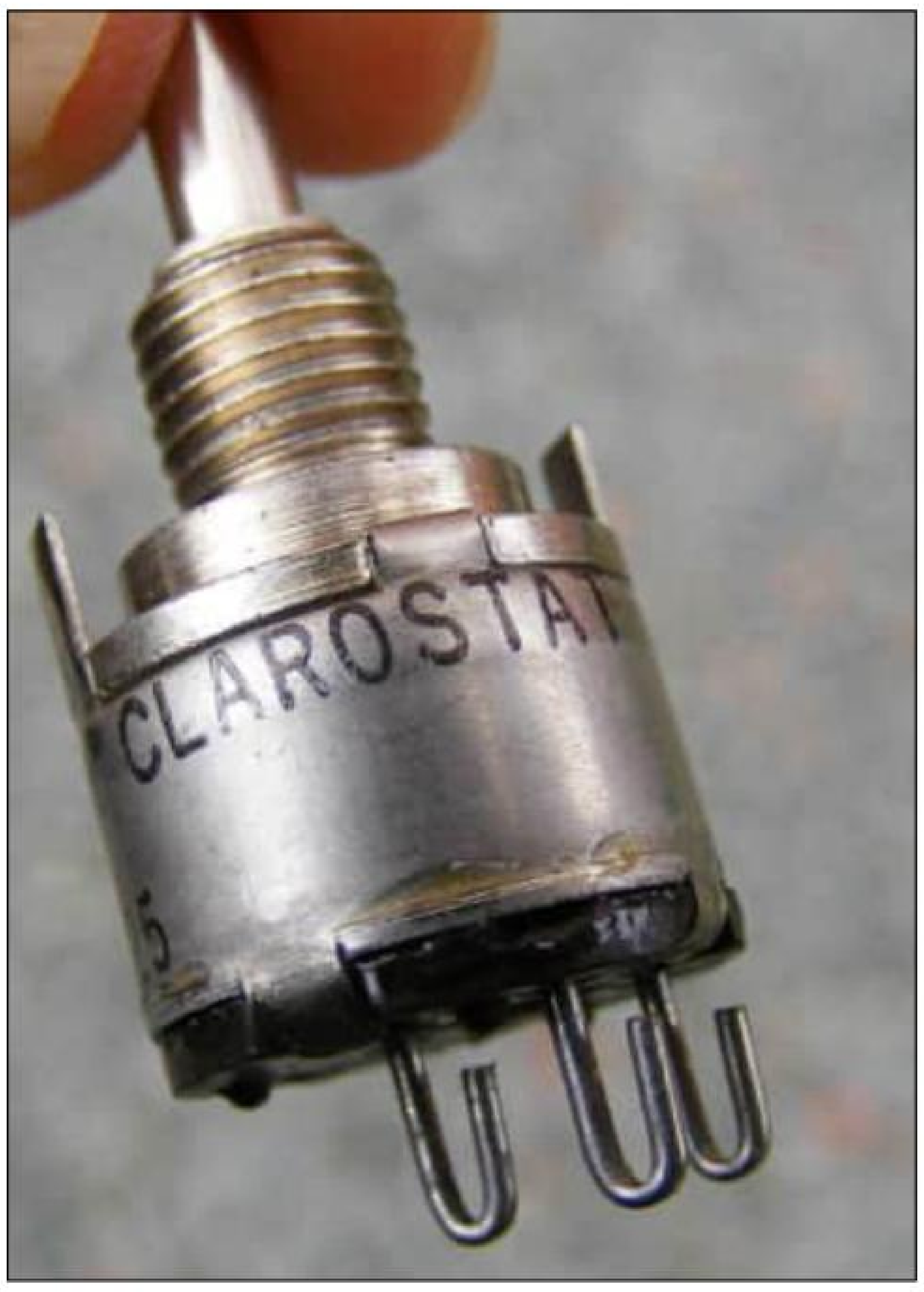


Fig.42. Clarostat RV6, a tiny mil-spec conductive plastic pot with a sealed stainless-steel case. I used these as the tuning control in my *Pocket Theremin* for many years.

Clarostat RV6 conductive plastic (now Honeywell) (100,000 RL)
Good, 10mm miniature, but expensive – see Fig.42. Has a dynamic noise resistance spec on its datasheet; in this case, ±1% of the track resistance. For the 50kΩ linear version I stock the expected variation would be ±500Ω.

Clarostat 388 modular series conductive-plastic (100,000 RL)
Top notch, used in SSL mixing consoles.

(Citec/TE connectivity MC1/U 25,000 RL)
Small size 0.5-inch, cermet high 2W rating. Well sealed, very solid. Used in Tornado jets. Similar in style to Plessey/Citec MH1 25,000 RL moulded-track version 0.25W rated. A bit inconsistent. The newer CLQ version has a rear epoxy end-seal to protect the pins from soldering damage. See Fig.43.

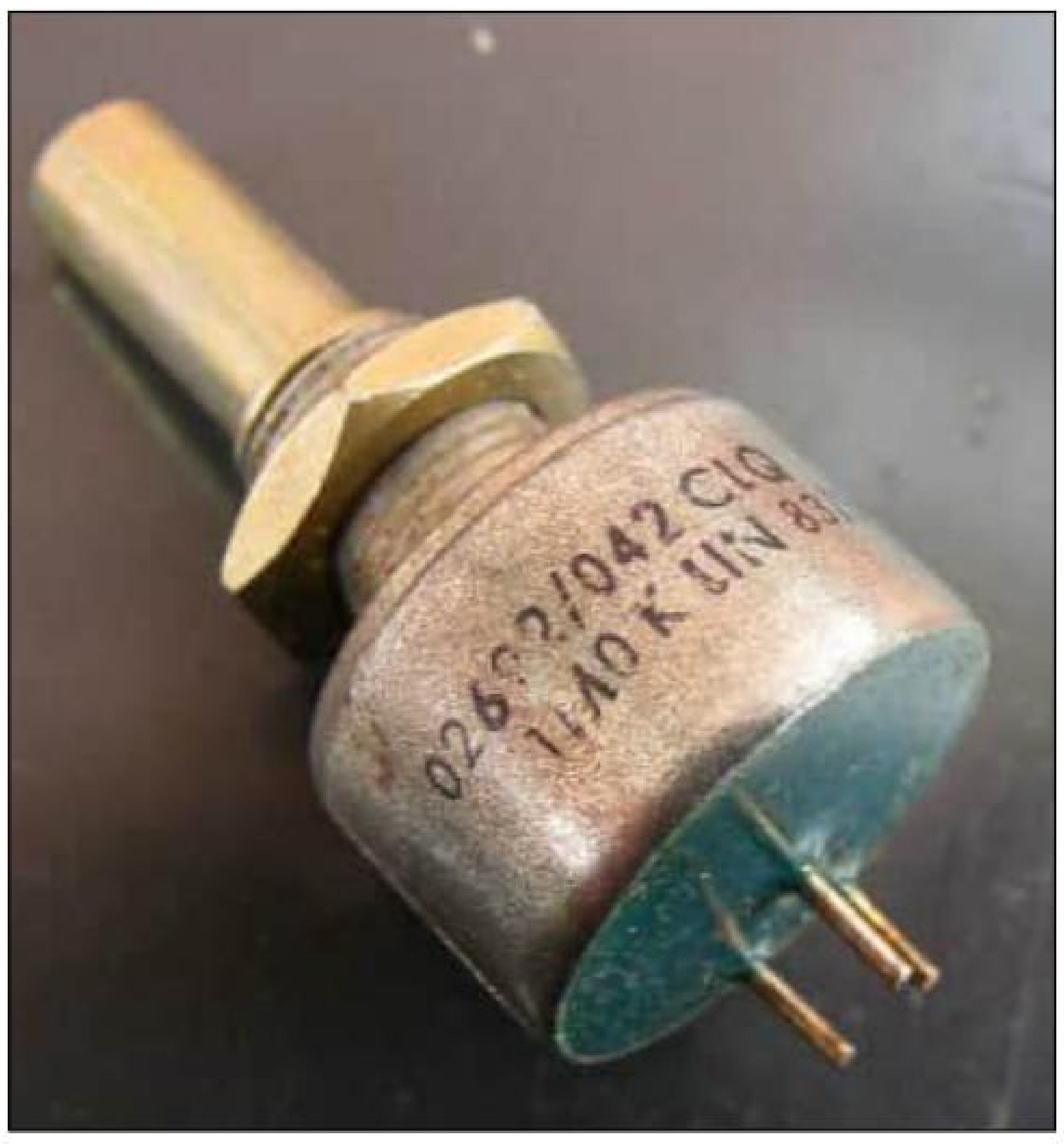


Fig.43. Citec/Plessey CLQ moulded track pot. Used in aviation equipment, very expensive, very tough. There is also a cermet MCU/1 version, but it's too noisy for volume controls.

Erie/ITT (100,000 RL)
Very old moulded-track design, very quiet for the era – see Fig.44.

Omeg (20,000 RL)
Early versions suffered from dissimilar slip ring metal, causing scratching after a year or so. These will have to be stripped out if found in equipment. Later versions with carbon polymer slip ring are much better.

Penny and Giles (>1,000,000 RL)
The very best, but stopped production recently. Audio Maintenance Ltd in Manchester still have a few stereo 104mm faders for £90.

Piher (10,000 RL)
Pretty rough, cheap with some awful plastic models.

Plessey E series (25,000 RL, but in reality more like 100,000 RL)
Very big, 2W-rated, solid moulded-track

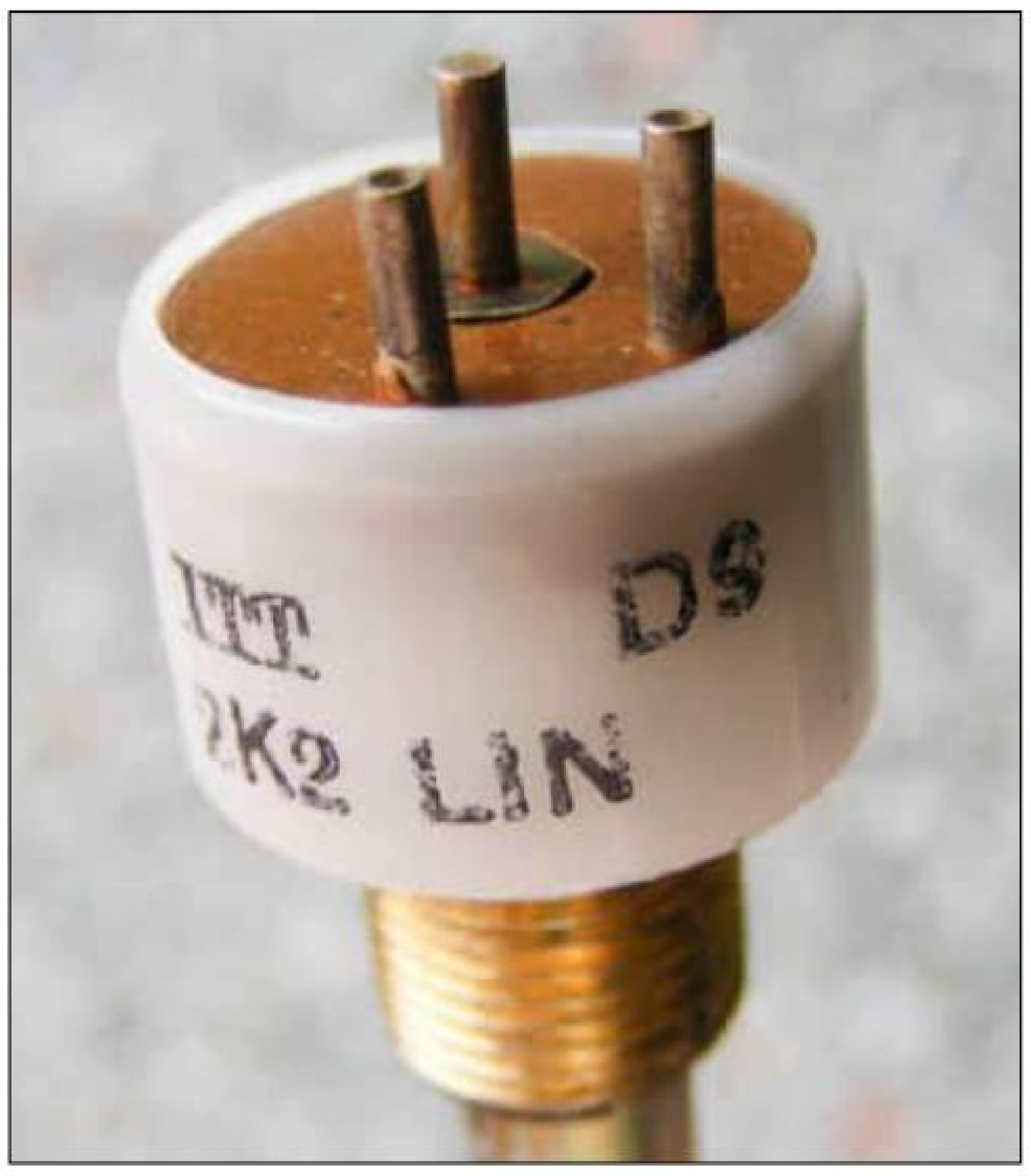


Fig.44. Erie moulded track pot, one of the few early ones designed for low rotational noise. I've used them in my guitar pedals with no problems.

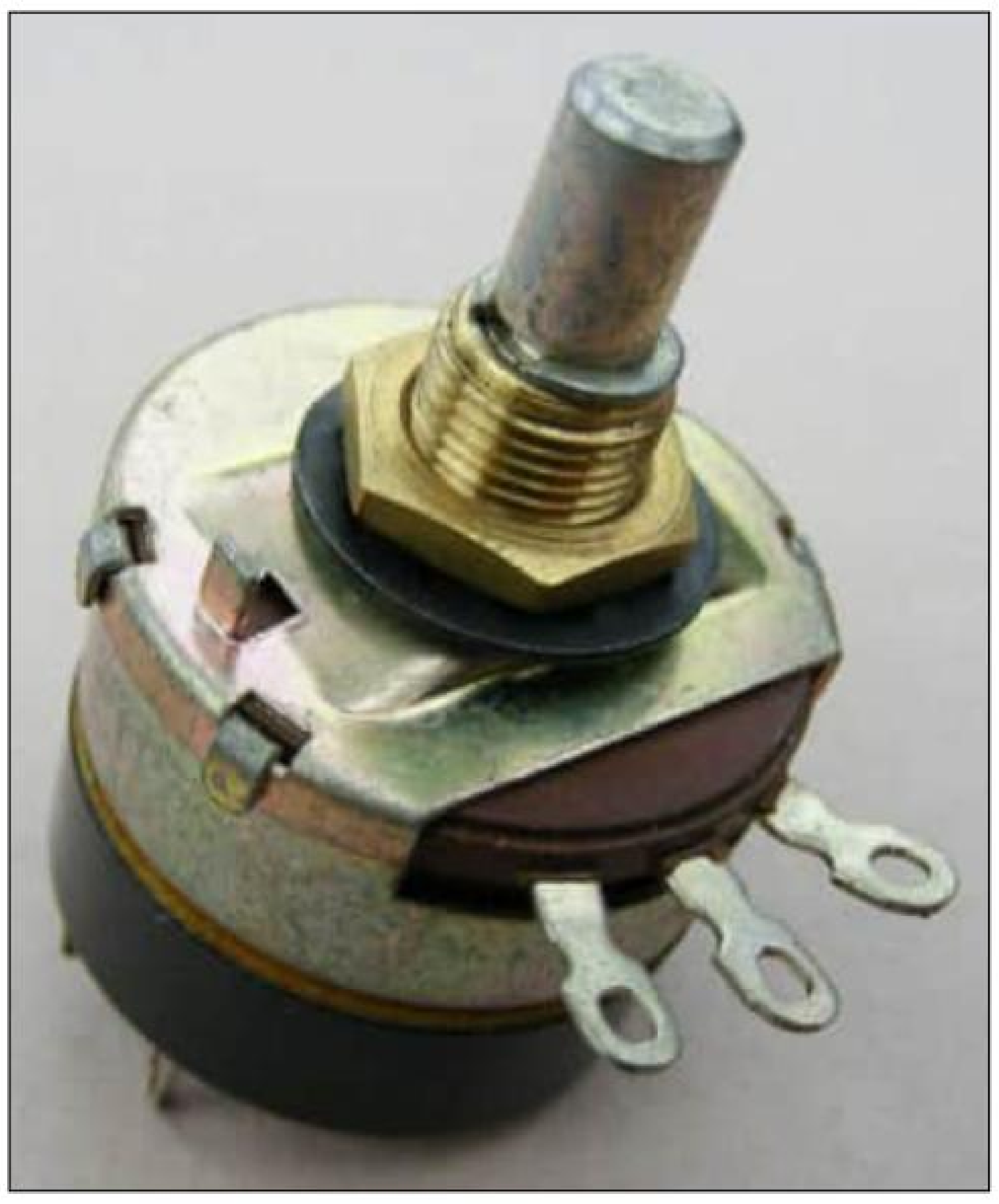


Fig.45. An old classic industrial moulded-track pot, the Plessey E series – they are almost indestructible.

and never seem to die. High torque, so they need a big knob, a bit of rotational play in the spindle. RS used to sell these under their own brand as 30mm, advertising them as having low rotational noise. However, given the roughness of the moulding it wasn't exceptional and had poor resistance tolerance. See Fig.45.

Ruwido (20,000 RL)
Good quality, small, sealed, sprayed-carbon instrumentation pots, popular with German audio engineers. Unusually, some are log, marked '+exp'. I used to use their faders which were physically strong. They were marked 'RUF', a rather unfortunate name since they were quite smooth. See Fig.46.

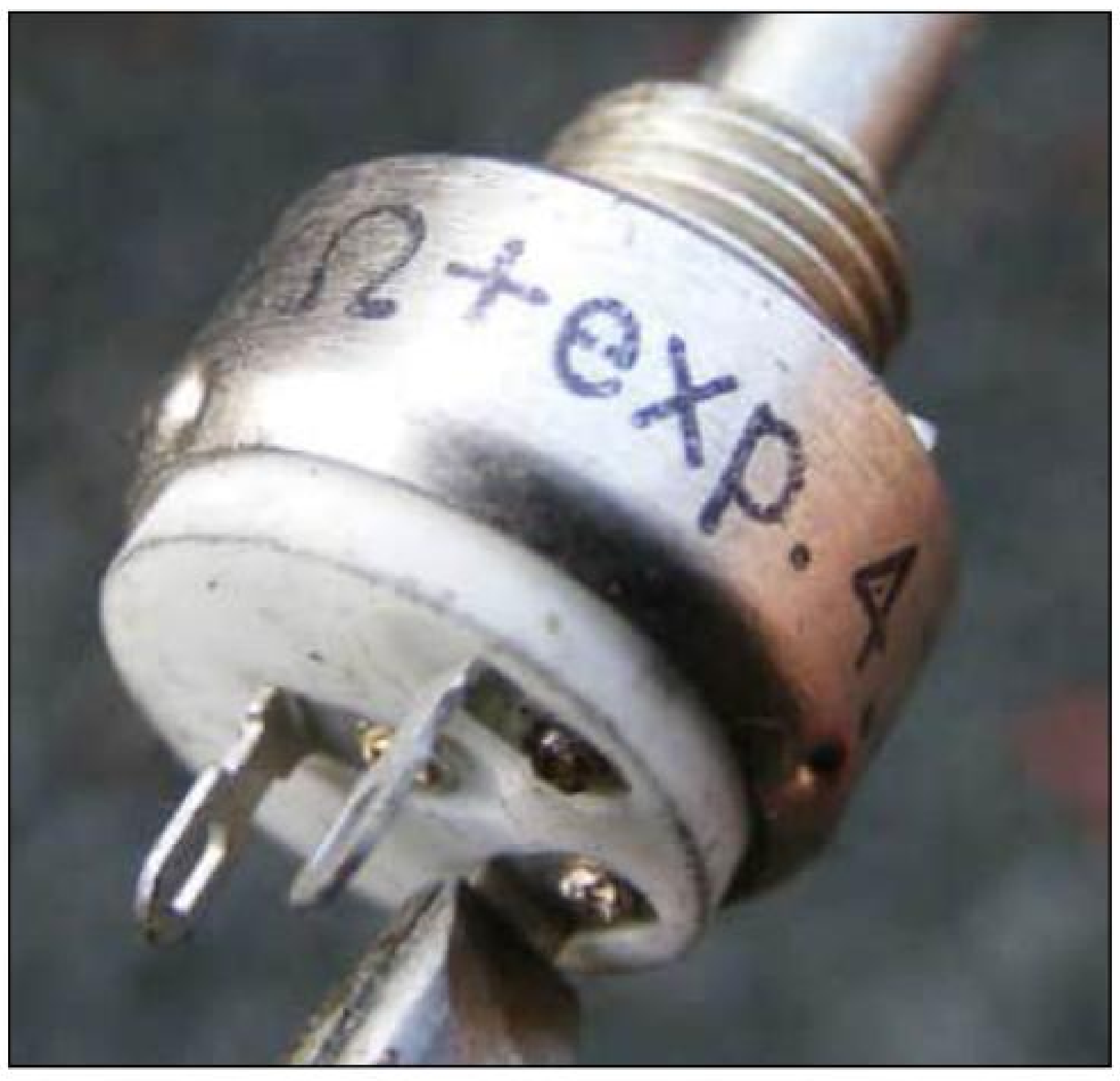


Fig.46. Very small Ruwido pot. Looks like a cermet due to its ceramic track substrate, but it actually has a log carbon film track with a carbon wiper.

Sfernice (100,000 RL)
Mainly cermet, but some are conductive plastic. Posh French units, very expensive, intended for instrumentation.

Spectrol 149 conductive plastic, small black cube shape (25,000 RL)
UK made, solid instrumentation grade.

CE Distribution in Arizona USA have a good selection of pots, mainly guitar related.

Just say 'no' to pots

One solution to all of the above problems is to simply avoid putting audio through pots and use a VCA (voltage-controlled amplifier) instead. I find the THAT VCA range of chips, such as the 2180xx series have sufficiently low distortion and noise for domestic Hi-Fi use. They also have an inherent logarithmic control characteristic. A stereo or six-channel surround volume control can be fed by a control voltage from a single linear pot smoothed by a capacitor. This increases the life of the pot several times, but for almost infinite life, a hall-effect sensor with a magnet on a shaft could be used. Where higher distortion and noise is acceptable (eg, in radios and synthesisers) transconductance amplifiers, such as the LM13600/700 or the better quality CA3280 can be used with a log control pot. Where stepped level control and a bit of 'zipper noise' is acceptable, there are digitally controlled switched resistive ladder-network chips available, such as the Dallas DS1802. These can use a rotary encoder, which can be optical, for a very long life. Of course, there's no scratching. All these solutions are relatively expensive, and much more complex than the humble pot. So, it seems the two main options today are 'software' volume controls or the simple, ancient potentiometer approach.

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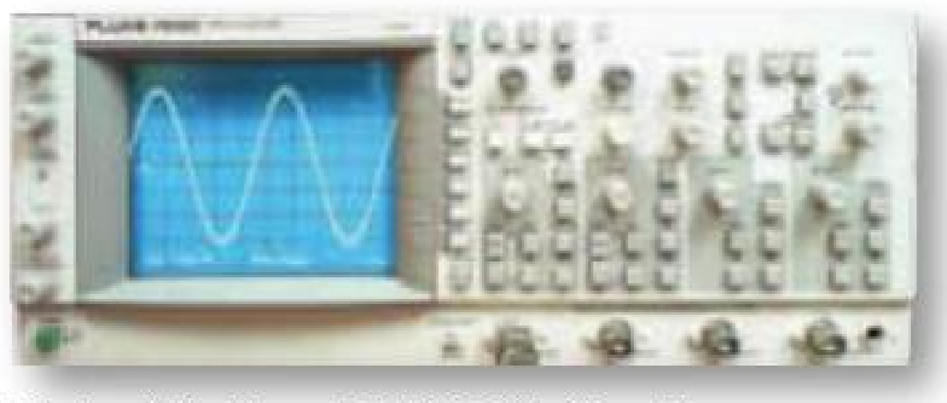

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Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

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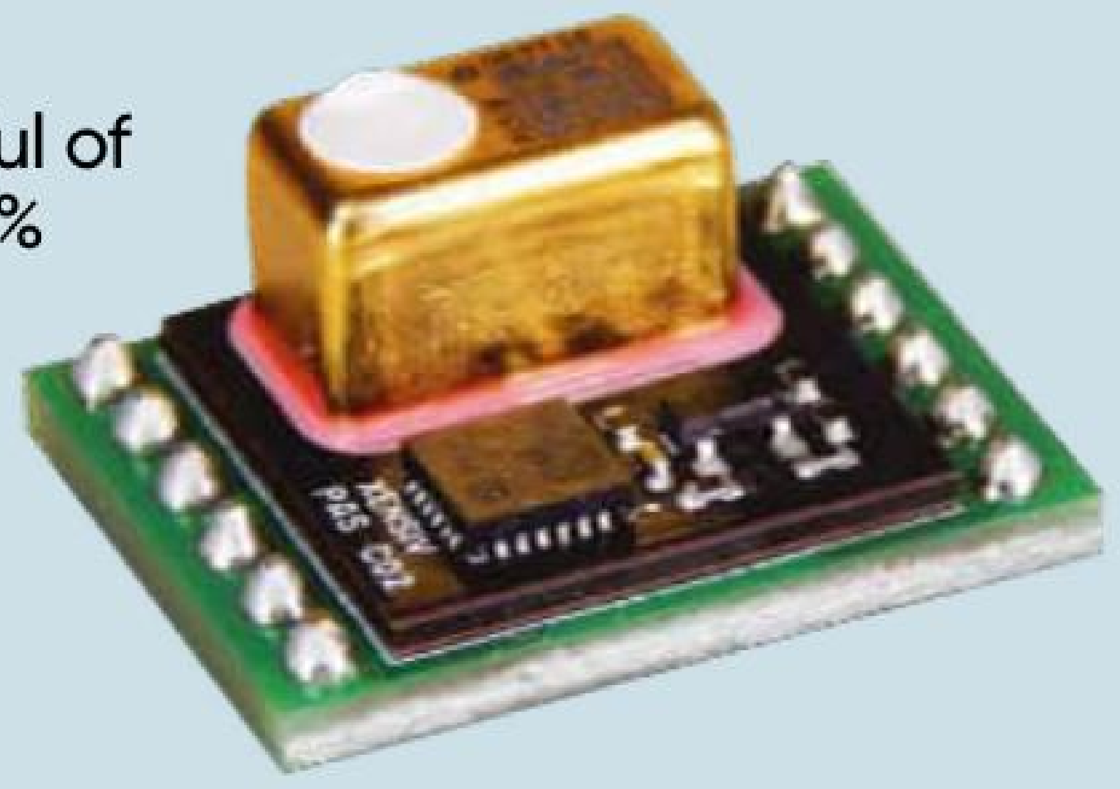
Next Month – in the August issue

Wide-Range Ohmmeter

This auto-ranging ohmmeter will measure just about any resistance – from a handful of milliohms to many megohms! It's accurate to better than ±1%; typically, close to ±0.1%

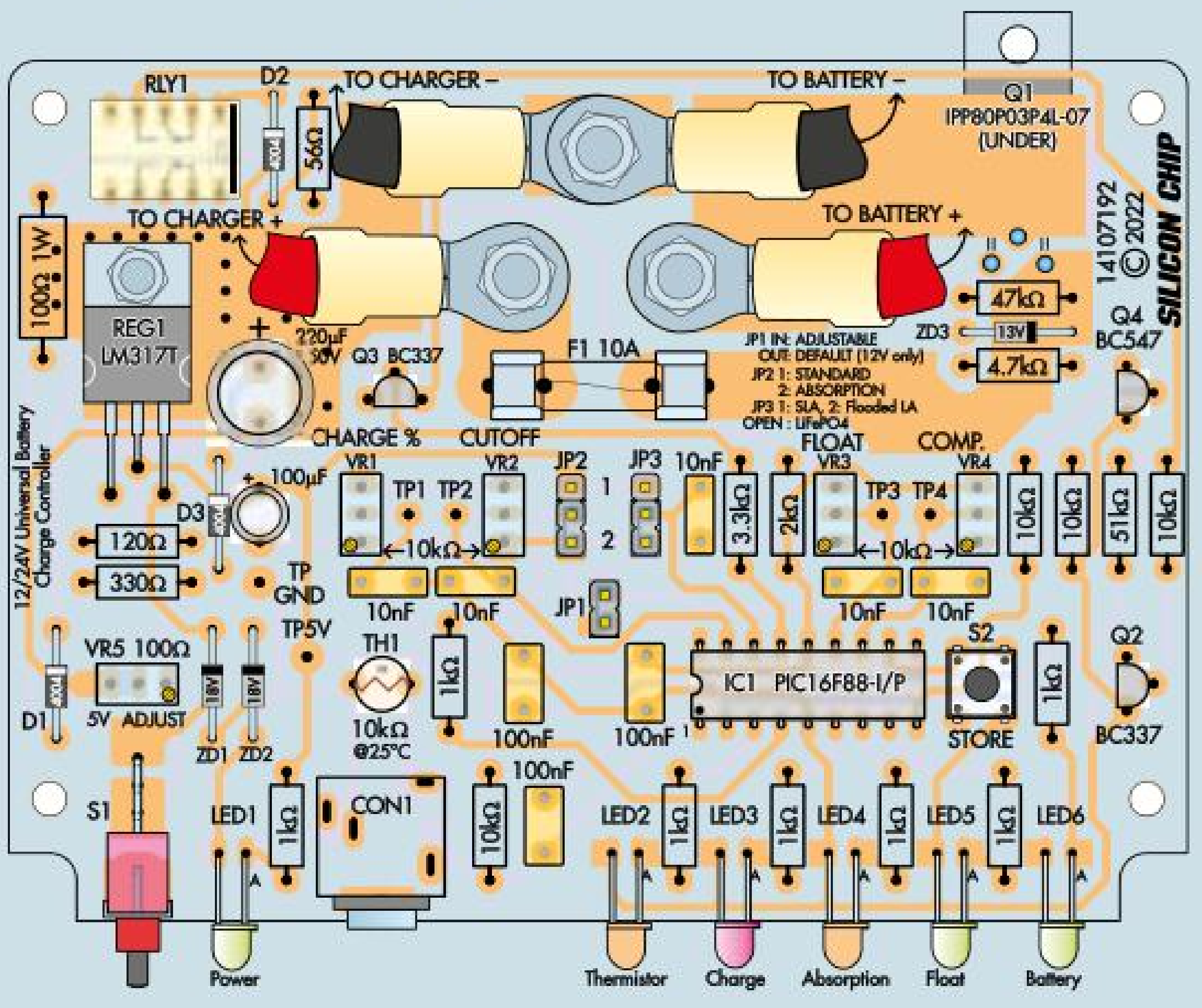


SPY-DER

This is a speech and web-controlled surveillance spider robot. It walks like a spider and acts as a spy using its camera, hence the name 'SPY-DER'. The best aspect of it is that you can make it yourself using some 3D-printed parts, a bunch of servos and some low-cost off-the-shelf electronic modules.

0-110dB RF Attenuator for Signal Generators

This attenuator is designed to accompany the AM/FM Signal Generator published in May 2023. However, you could combine it with just about any signal generator to provide easy output level adjustment over a wide range.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Universal Battery Charge Controller – update!

We have made some changes to our Universal Battery Charge Controller published in December 2020 so that it can be built using parts that are actually available. The changes are modest, and the total cost for parts has been reduced.

PAS CO2 Air Quality Sensor Module

Continuing our examination of low-cost air quality sensors, next month we take a look at the Infineon XENSIV PAS CO2 mini-board.

PLUS!

All your favourite regular columns from *Audio Out, Cool Beans and Circuit Surgery*, to *Make it with Micromite, Techno Talk* and *Net Work*.

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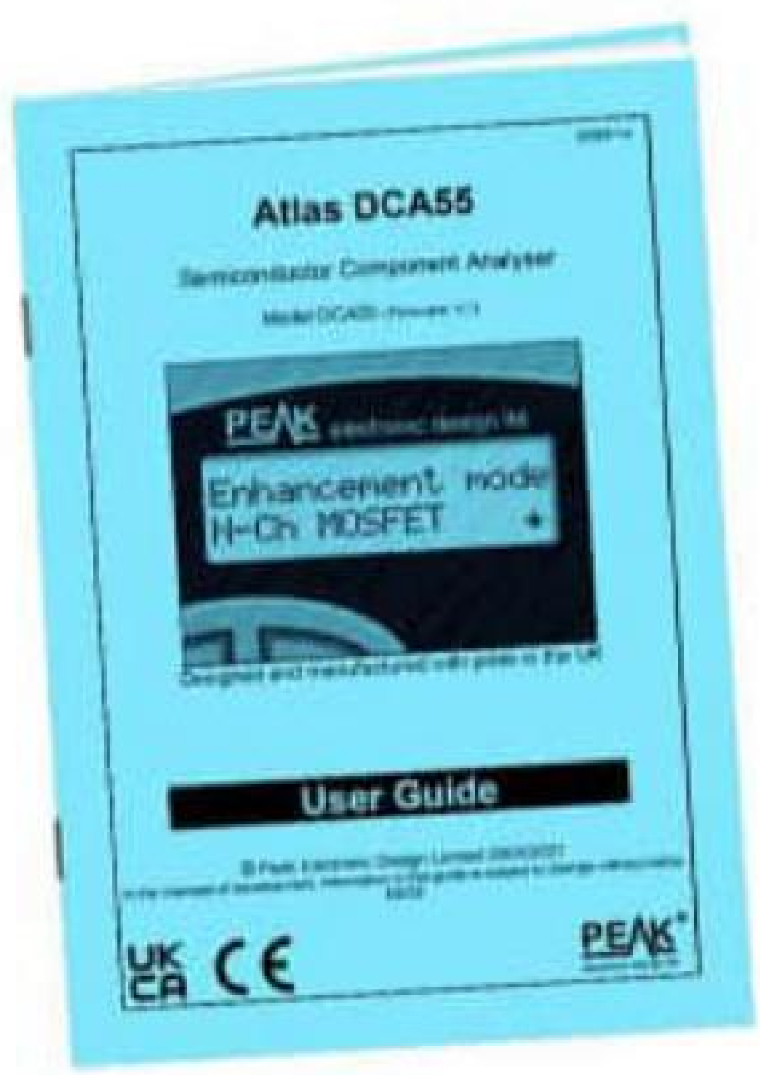

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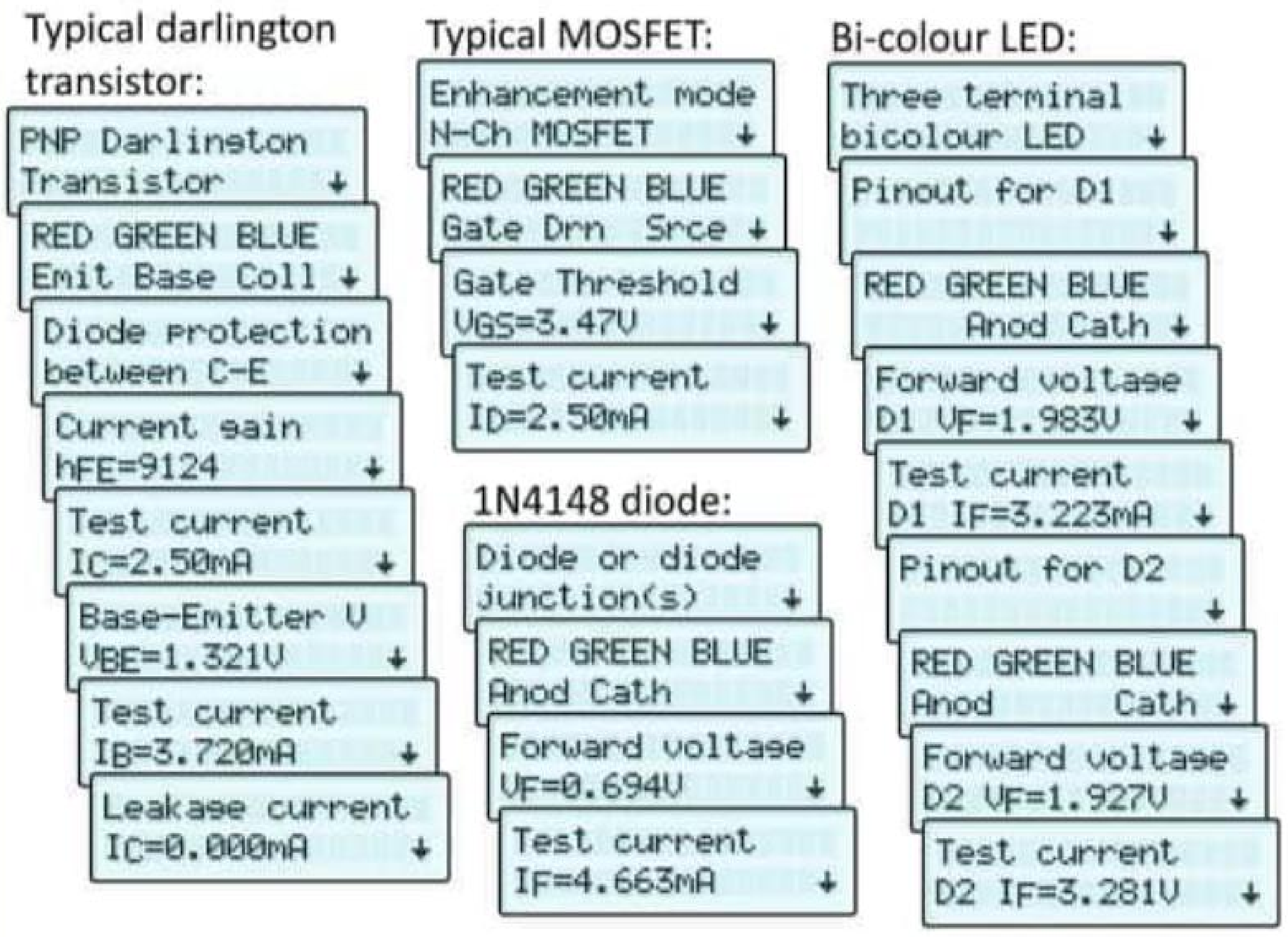

Typical darlington transistor:

PNP Darlington
Transistor ↓

RED GREEN BLUE
Emit Base Coll ↓

Diode protection
between C-E ↓

Current gain
hFE=9124 ↓

Test current
IC=2.50mA ↓

Base-Emitter V
VBE=1.321V ↓

Test current
IB=3.720mA ↓

Leakage current
IC=0.000mA ↓

Typical MOSFET:

Enhancement mode
N-Ch MOSFET ↓

RED GREEN BLUE
Gate Drn Srce ↓

Gate Threshold
VGS=3.47V ↓

Test current
ID=2.50mA ↓

1N4148 diode:

Diode or diode
junction(s) ↓

RED GREEN BLUE
Anod Cath ↓

Forward voltage
VF=0.694V ↓

Test current
IF=4.663mA ↓

Bi-colour LED:

Three terminal
bicolour LED ↓

Pinout for D1
↓

RED GREEN BLUE
Anod Cath ↓

Forward voltage
D1 VF=1.983V ↓

Test current
D1 IF=3.223mA ↓

Pinout for D2
↓

RED GREEN BLUE
Anod Cath ↓

Forward voltage
D2 VF=1.927V ↓

Test current
D2 IF=3.281V ↓

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